F.M. Radio Servicing Handbook

2nd Edition

Gordon J. King

Newnes-Butterworths

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by Gordon J. King

THE PRACTICAL AERIAL HANDBOOK

RADIO AND AUDIO SERVICING HANDBOOK

RADIO AND TELEVISION TEST INSTRUMENTS

TELEVISION SERVICING HANDBOOK

THE HI-FI AND TAPE RECORDER HANDBOOK

SERVICING WITH THE OSCILLOSCOPE

RAPID SERVICING OF TRANSISTOR EQUIPMENT

In preparation

COLOUR TELEVISION SERVICING

F.M. RADIO SERVICING HANDBOOK

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LONDON

NEWNES—BUTTERWORTHS

THE BUTTERWORTH GROUP

ENGLAND	BUTTERWORTH AND CO. (PUBLISHERS) LTD. LONDON: 88 Kingsway, WC2B 6AB
AUSTRALIA	BUTTERWORTH AND CO. (AUSTRALIA) LTD. SYDNEY: 20 Loftus Street MELBOURNE: 473 Bourke Street BRISBANE: 240 Queen Street
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SOUTH AFRICA	BUTTERWORTH AND CO. (SOUTH AFRICA) (PTY.) LTD. DURBAN: 33/35 Beach Grove

First published 1957
Reprinted 1958
Reprinted 1959
Reprinted 1960
Reprinted 1967
Second Edition 1970

Published in 1970 by Newnes-Butterworths,
an imprint of the Butterworth Group

ISBN 0 408 00023 6

PHOTOTYPESET BY FILMTYPE SERVICES LTD., SCARBOROUGH, ENGLAND
PRINTED BY THE GARDEN CITY PRESS LIMITED, LETCHWORTH, HERTFORDSHIRE

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Preface

I WAS GLAD to have had this opportunity of completely revising this rather important f.m. book of my series to coincide with the "coming of age", so to speak, of the f.m. system of broadcasting. Radio listeners are now well appreciating the better quality of reproduction made possible only by the f.m. system, and the limitations of the small transistor radio in this respect are better understood.

True some hi-fi types still complain—often rightly so—of the relatively poor transmission quality of certain programmes when they are conveyed over extended cable links to f.m. stations well removed from the audio sources; but the BBC is actively engaged in establishing even higher standards of quality, and as more use is made of super-high-frequency (s.h.f.) radio links to carry two-channel stereo to the main transmitters, the quality will improve and the complaints diminish.

The advent of f.m.-stereo is, indeed, a great hi-fi landmark which occurred between the first and this second edition—a matter of twelve years. Although stereo has failed to spread across the country as rapidly as some enthusiasts and music lovers had hoped, it might well be that before the next edition of this book is due most parts of the U.K. will be within range of at least one stereo-encoded station.

Improved sensitivity and noise performance of the new generation of f.m. receivers and tuners brought about by low-noise field-effect transistors and integrated circuits (i.c.s) of the space age are making it possible to receive stereo-encoded transmissions well in advance of the accepted v.h.f. range—albeit, somewhat irregularly as dictated by the prevailing tropospheric conditions. Down here in the heart of South Devon—200 miles from Wrotham and 160-odd miles from its booster station at Oxford—I have been able to enjoy stereo broadcasts even though they appear with more background noise than I would wish for most of the time owing to the extended range of reception.

This new edition deals fully with the stereo system and contains information on the latest f.e.t. front-ends and integrated-circuit i.f. channels. I have scrapped quite a bit of the original material and added new, including a new Chapter 7 on the stereo system and information on transistorized equipment of all kinds. Hence a reader with a copy of the first edition will not find that he has in essence a duplicate should he decide to invest in a copy of this new edition!

To conclude I would like to thank the many readers of the first edition, especially those whom have written to me, also the manufacturers of f.m. equipment and solid-state devices for allowing me to draw upon them for information, without whose co-operation this book could not have been written. My thanks also go to the BBC and to all my friends in the radio, TV and hi-fi world; and to my company for permission to publish certain photographs.

Brixham

G.J.K.

CHAPTER 1

The F.M. Signal

FREQUENCY MODULATION (f.m. for short) was put forward as a system of communication in the very early days of radio. The basic idea was germinated before the 1920s, but in those days it was viewed with considerable scepticism. The system was given more serious consideration in 1925 by the late Major E. H. Armstrong, the true f.m. pioneer, who advocated that its adoption would provide enhanced freedom from interference of all kinds, for even in those days congestion was becoming apparent in the normal broadcast bands.

In the early work on the system it was thought feasible to transmit the audio intelligence by causing the carrier frequency to fluctuate by only a few hundred Hertz at a rate equal to the modulation frequency, thereby permitting stations to be spaced more closely than the 9 or 10 kHz spacing demanded when using amplitude modulation (a.m.).

This reasoning was disproved when it was later revealed mathematically that a frequency-modulated carrier would create multiple sidebands which would tend to "splash" into an adjacent broadcast channel and thus cause interference on a scale far larger than that already existing due to sideband splashing of the a.m. signals.

The advantages attendant on the use of the f.m. system were not fully realized until engineers became more knowledgeable about the possibilities offered by transmission and reception in the very-high-frequency (v.h.f.) bands.

It was not until 1936, following Armstrong's classic paper "A method of reducing disturbances in radio signal signalling by a system of frequency modulation", when the interference problem seemed like becoming insoluble, that the system was explored with fresh zeal. It was progressively developed experimentally in the USA some years before the war and was extensively used for communication purposes during the war.

After the war it was adopted in America for domestic broadcasting while in Europe the system, as applied to domestic broadcasting, was viewed controversially. It was argued that the a.m. system would be endowed with all the advantages claimed for the f.m. system, given the same conditions and frequency spectrum, and that efficient noise-suppression circuits added to high-quality a.m. receivers would balance the interference-free claims of the f.m. system. The merits and demerits of the system finally became the subject of a somewhat emotional transatlantic debate.

At this time Great Britain seriously began to consider using the system for domestic broadcasting. The problem was approached from the practical aspect, and both a.m. and f.m. test transmissions were provided in the v.h.f. band from Wrotham in Kent in a final endeavour to settle which system would be better suited to British requirements. The f.m. system was finally decided on and it has now developed into a nation-wide means of high-quality broadcasting.

Curiously, during the time that Britain was debating the use of f.m., its popularity in America started to decline; but the interest in high-quality reproduction eventually resulted in a revival. The f.m. system, however, is universally adopted for local communication services and forms the standard of sound-channel modulation both in American and Continental television systems.

Accelerated by local broadcasting and the BBC v.h.f./f.m. national programmes, the demand for f.m. equipment is now showing signs of a significant increase in contrast to a period of stagnation for a few years following the publication of the first edition of this book. Many thousands of f.m. tuners are currently employed in partnership with domestic hi-fi sound systems and so-called "music compacts" are also adopting the system, and f.m. is even finding its way into car radios! Moreover, it is the standard system for the sound channel of 625-line TV. Hitherto endeavours to obtain high-quality reception were limited to the adverse factors associated with the a.m. system of broadcasting, the chief one of which is the required quashing of the overall a.f. passband as a means of eliminating whistles and cross-talk caused by interference originating from an unwanted station occupying an adjacent channel.

The f.m. system also makes it possible to transmit two separate a.f. channels on a common carrier by a method of multiplexing—explained in Chapter 7—to provide two-channel stereo reception.

The v.h.f./f.m. system provides the very maximum of channel width and allows transmission and reception of the full range of audio frequencies without the worry of interference. It is true, of course, that the a.m. system, given the same operating frequency (v.h.f.) and bandwidth, would also be capable of providing a very high-quality broadcast service, but with f.m. there are other desirable factors which are considered to contribute to its advantages over the a.m. system.

For example, it is interesting to note that with wide-band f.m. a spurious signal, actually within the channel of the required station, would not cause interference unless its strength was in excess of half that of the required signal, depending on the "capture ratio", the more powerful signal always predominates, and the weaker one is pushed right into the background so that it is not heard.

It is remarkable that in order to achieve the same degree of success on a.m. the wanted station would have to be about 25 times the strength of the unwanted one. Moreover, the f.m. transmitter can be operated continuously at peak power as the carrier is changed only in frequency and not in amplitude. Apart from assisting with the problem of interference, this factor provides a two-to-one better signal for a given capital expenditure on the transmitter over the a.m. system. This is because the unmodulated carrier of the a.m. station can only have half the peak amplitude it reaches at 100-per-cent modulation.

CHARACTER OF THE F.M. SIGNAL

Any system of modulation used to transmit audio intelligence must be capable of conveying two characteristics, frequency and amplitude; these are required to provide the frequency and loudness variations of the original sound. The basic principle of the f.m. system is that the transmitting carrier wave fluctuates on either side of the nominal frequency at a rate depending on the frequency of the applied audio intelligence or modulation signal, while the magnitude of the frequency fluctuations is governed by the amplitude or loudness of the modulation. The *frequency* of the carrier wave is thus modulated by the audio intelligence, while the amplitude of the carrier wave remains essentially unaltered as the result of the modulation.

In order to obtain a better understanding of the frequency-modulated carrier, let us refresh our memories on the basic principle of a.m. It will be recalled that the *amplitude* of the carrier wave varies in direct sympathy with the wave pattern of the modulation it is to carry, and that the frequency of the a.m. carrier is virtually unaffected by the modulation signal. The frequency of the modulation gives rise to a wave pattern on the carrier wave, usually referred to as the modulation envelope, while the amplitude or loudness of the applied modulation governs the extent or depth to which the modulation wave pattern penetrates into the carrier wave.

The principles of the two systems can be assimilated more readily from diagrams of the various waveforms. At Fig. 1.1 is depicted the a.m. case. At (b) is shown an audio-frequency (a.f.) sine wave, representing a constant tone, such

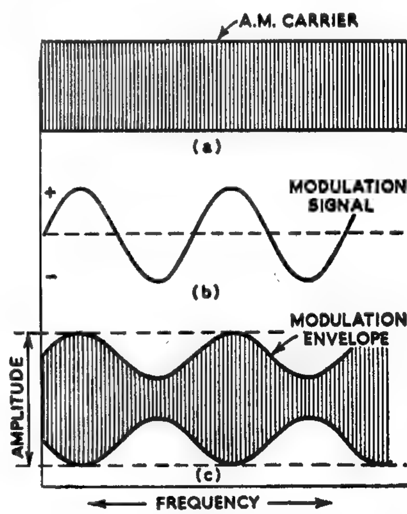


FIG. 1.1. *(a) Diagrammatic representation of amplitude modulation; (b) an audio-frequency sine wave as employed to modulate the carrier wave; (c) the waveform of the a.m. signal.*

as a test signal, which might well be used to modulate the radio-frequency (r.f.) carrier wave at (a). The resulting amplitude-modulated carrier would thus appear in waveform as at (c).

Let it be supposed that the frequency of the carrier wave is 5MHz and it is required to carry an audio signal of 500 Hz. The audio tone is applied at the transmitter in such a way that the carrier remains unaffected from the frequency aspect—it continues at 5 MHz—but its amplitude increases during the time the

modulation waveform swings positive, and decreases during the time it swings negative (Fig. 1.1).

The carrier wave thus develops a modulation envelope which follows the wave pattern of the 500 Hz modulation signal. In this form the carrier wave is radiated from the transmitting aerials and so carries the full character of the modulation, and this can readily be extracted and converted back into the 500 Hz tone at the receiver.

A representation of the f.m. case is given in Fig. 1.2, where at (b) is shown the a.f. modulation signal which is used to modulate the r.f. carrier wave at (a). The resulting frequency-modulated carrier appears as at (c).

Let us again suppose that the carrier frequency is 5 MHz and the modulation 500 Hz. Here, the modulation is applied at the transmitter in such a way that

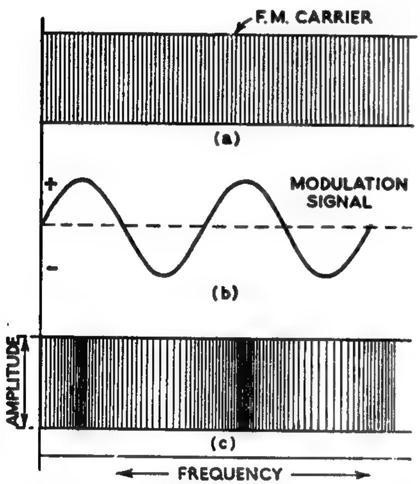


FIG. 1.2. *At (b) is shown the signal used to modulate the carrier wave at (a); (c) the resulting f.m. signal waveform. Straight lines have been used in (a) and (c) to represent lines of sinusoidal character.*

the amplitude of the carrier remains constant but it fluctuates from its nominal unmodulated value at a rate governed by the frequency of the modulation waveform; the frequency of the carrier increases during the time the modulation waveform swings positive and decreases during the time it swings negative, as in Fig. 1.2.

The extent by which the carrier fluctuates from its unmodulated value is not related to the frequency of modulation. For example, the 500 Hz modulation will cause the carrier frequency to fluctuate to and fro above and below 5 MHz at five hundred times per second. The extent of the fluctuation depends on the loudness or amplitude of the modulation signal; the extent of fluctuation will be the same with modulation signals of all frequencies when their amplitudes are equal, excluding the effect of pre-emphasis (see later).

The degree of fluctuation of the f.m. carrier is synonymous to the depth of modulation in the a.m. case. In both cases it is the amplitude or loudness of the modulation signal which decides how deeply the carrier wave is modulated.

Let us go back to the a.m. case and refer to Fig. 1.1 again. Here the a.m. waveform at (c) is representative of something less than 100-per-cent modulation. If the amplitude of the modulation signal at (b) were larger and of the right

value to cause the amplitude of the carrier wave to fall exactly to zero at the negative peaks of the modulation, then the carrier would be modulated to its maximum, referred to as 100-per-cent modulation.

It is clear that if the modulation signal had been of greater amplitude than that required to provide 100-per-cent modulation the carrier would have fallen to zero and remained there over a period of the actual modulation signal, which would result in severe distortion.

Reverting to the hypothetical consideration of a 5 MHz carrier and 500 Hz modulation, and assuming 100-per-cent modulation, the carrier wave would attain an instantaneous amplitude twice as large as its unmodulated amplitude at a rate of 500 times per second and would fall to zero half-way between each of the maximum peaks. It is of interest to note that if (a) and (c) in Fig. 1.1 were drawn to scale, the average amplitude of (c) would equal the amplitude of (a) at all levels of modulation within the 100-per-cent limit.

Consider how the modulation amplitude affects the f.m. case. The degree of fluctuation above and below the nominal frequency of the carrier is known as the *deviation frequency* and is measured in kHz. Since both modulation and deviation are measured in kHz, care must be taken not to get the two confused.

The maximum deviation as related to 100-per-cent modulation is not wholly limited as with a.m. but, within stipulated limits, can be chosen to suit the specific purpose of the f.m. system as a whole. Later it will be seen how the deviation is somewhat tied up with the question of interference and the audio bandwidth of the f.m. system. For the present it is sufficient to say, roughly, that the minimum deviation limit is governed by the highest modulation frequency it is required for the system to carry.

At one time it was incorrectly postulated that a high-quality audio signal could be carried by f.m. within a channel width corresponding to the deviation

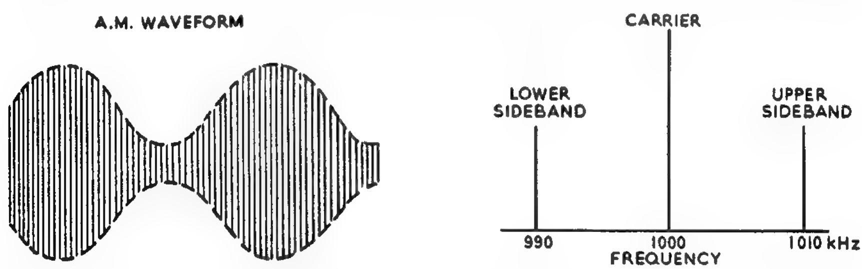


FIG. 1.3. *An a.m. waveform can be resolved into three parts, the carrier wave, an upper sideband and a lower sideband.*

frequency. It was argued that modulation signals of the very highest audio frequencies could be used to modulate a carrier whose frequency was being deviated by only a few hundred Hertz, and thus, it was thought, considerably conserve spectrum space, but then engineers were not fully conversant with the problems of sidebands.

Before discussing further the question of sidebands, it should be known that the maximum deviation frequency of the existing BBC frequency-modulated transmissions is 75 kHz. This means that for 100-per-cent modulation the carrier

is deviated 75 kHz above and below its nominal frequency, and 37·5 kHz for 50-per-cent modulation and so on, remembering that the overall frequency swing represents twice the deviation value, or 150 kHz for 100-per-cent modulation, since the deviation frequency is really plus and minus 75 kHz.

Since it was intimated that the minimum deviation limit is set by the highest modulation frequency, one might ask why a deviation of 75 kHz is used when this is well above the limit of even the highest audio frequency. The answer to this will be given later when the factors relating to interference are explored. For the present it should be realized that the modern f.m. system offers no restrictions to modulation signals of any frequency within the audio spectrum.

DEALING WITH SIDEBANDS

As an aid to the understanding of the f.m. case, we will continue for a while to make comparative references to the a.m. system. From the aspect of sidebands, such comparisons can prove extremely useful.

With amplitude modulation, the effect of applying modulation of a constant frequency is virtually the same as having a carrier wave of constant frequency and amplitude and adding two additional component frequencies known as sidebands. The frequency of the sidebands is directly related to the frequency of the modulation signal, while their amplitude is governed by the depth of the modulation. Two sidebands, each of half the amplitude of the carrier wave, are created when the carrier is modulated to a depth of 100 per cent.

Supposing, for example, a carrier wave at 1,000 kHz is modulated at 10 kHz, then the combined a.m. waveform could be resolved into three parts: (1) the carrier, (2) the upper sideband at 1,010 kHz, and (3) the lower sideband at 990 kHz. A composite signal would be obtained extending from 1,010 kHz to 990 kHz, or over a range of 20 kHz, which is twice the modulation frequency. This is shown in Fig. 1.3.

With f.m., things are not quite like this; although sidebands are produced they are a little more involved. From what has been seen so far, it may appear that, with a carrier frequency of, say, 90 MHz which is deviated plus and minus

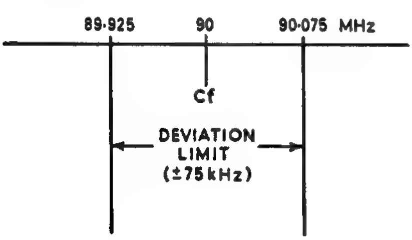


FIG. 1.4. *Illustrating deviation frequency. Here, the carrier frequency, Cf, is 90 MHz and the deviation limit plus and minus 75 kHz*

75 kHz for full modulation, the overall frequency requirements of the resulting f.m. signal could well be occupied within the deviation limit, that is within 89·925 and 90·075 MHz, as shown in Fig. 1.4.

It may be reasoned that since the highest modulation frequency needed for high-quality sound is unlikely to exceed 15 kHz, any sidebands due to such a frequency would fall within the deviation limit anyway. This would be true from the a.m. point of view, where the composite signal made up of the carrier and

two sidebands would extend only 15 kHz above and below the carrier frequency and virtually become lost within the 150 kHz spectrum corresponding to the f.m. deviation limit. It is not possible to consider the two systems in this light, because with f.m. a modulated carrier creates a *series* of pairs of sidebands which are separated from the carrier by one, two, three and so on times the modulation frequency, as depicted in Fig. 1.5. Just a single modulation signal of constant

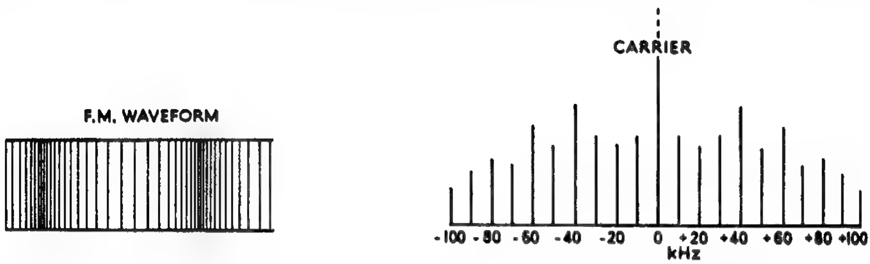


FIG. 1.5. *With f.m., a modulated carrier creates a series of pairs of sidebands which are separated from the carrier by one, two, three and so on times the modulation frequency. As the high-order sidebands diminish in amplitude, a bandwidth of 200 kHz will embrace all those of appreciable amplitude.*

frequency, in fact, gives rise to a series of pairs of sidebands, instead of only one pair as with a.m. From this, the complex nature of the sidebands resulting from modulation signals corresponding to a full orchestra can be imagined. For present purposes there is no need to become too involved in theoretical considerations of the sidebands, however, it being sufficient to know how they affect the make-up and performance of the receiver generally.

It was seen earlier that the deviation frequency must not be less than the highest modulation frequency it is required to transmit. If it is assumed, for the time being, that the deviation does, in fact, correspond to the highest modulation frequency, which will be considered as 15 kHz, then this means swinging the carrier over 30 kHz (plus and minus 15 kHz). In terms of bandwidth, this is the same as required by an a.m. system modulated in the same way. With a.m., however, the signal componenets are complete within 30 kHz. With f.m. the signal components, comprising the multiple sidebands, spread out considerably beyond this limit.

Fortunately, as the sidebands become farther displaced from the carrier their amplitude tails off, and after a while become sufficiently small to be neglected. As with amplitude modulation, f.m. sidebands depend upon the modulation frequency; if this is put at 15 kHz, being the upper audio limit, and the deviation is 75 kHz, for 100-per-cent modulation, then a bandwidth in the region of 200 kHz will embrace all sidebands greater than 1½ per cent of the undmodulated carrier. Nevertheless, sidebands of appreciable amplitude will be present up to about 50 or 60 kHz above and below that of the carrier.

It has already been intimated that the standard accepted deviation of 75 kHz is not truly to enhance the modulation frequency range but to ease interference problems. Indeed, a high-quality f.m. system could readily be accommodated in a 50 kHz channel employing a maximum deviation considerably below this

frequency, but then one of the main advantages attendant on the use of the f.m. system would not be fully realized.

The ratio between the amount that the carrier wave fluctuates above and below its nominal frequency (really the depth of modulation at any given instant) and the frequency of modulation is termed the *modulation index* (M). It refers to the sideband structure of an f.m. transmission at any given instant during a programme and is continually varying. If M is very small, then the f.m. signal components might well only extend to twice the modulation frequency, the same as with a.m., but as M is made larger so the sideband components of the signal extend proportionally. M is expressed as the deviation frequency divided by the modulation frequency.

Clearly, the extended f.m. sidebands need to have sufficient channel width in which to develop fully without overlapping into another channel. This, therefore, makes it clear that the function of the frequency modulation of a carrier wave does not in itself resolve the problem of interference. The problem of interference due to sideband splash might well, in fact, be considerably aggravated. The long-wave broadcast band, for instance, would scarcely be wide enough over its entire range to cater for a single f.m. channel, while, in favourable circumstances, the medium-wave broadcast band would hold about two normal channels. It can well be realized why early protagonists of f.m. were forced to put aside their experiments which were aimed at relieving the increasing congestion on the medium-wave band by using a carrier of narrow deviation, when they suddenly became enlightened concerning the production of multiple f.m. sidebands, bearing in mind that little was known of v.h.f. transmissions in those days.

It is obvious that f.m. channels must be accommodated in the v.h.f. regions where there is plenty of room for full expansion of the sidebands. The f.m. channels used by the BBC are 200 kHz wide, and thus give complete immunity from sideband interference due to stations operating on adjacent channels, even when the f.m. signals swing to full modulation and carry the very highest audio frequencies.

Band II is a part of the v.h.f. spectrum extending from about 88 to 108 MHz. It is divided into 200 kHz f.m. channels. Table 1.1 shows the national stations of the BBC and Table 1.2 shows the local stations.

The remaining v.h.f. bands are: Band I extending from about 40 to 70 MHz, currently divided into five 5 MHz 405-line TV channels, and Band III extending from about 170 to 216 MHz, currenly divided into eight 5 MHz 405-line TV channels. Bands IV and V fall into the u.h.f. spectrum and carry 625-line monochrome and colour programmes in 8 MHz channels.

PRODUCTION OF A FREQUENCY-MODULATED SIGNAL

Although concerned mainly with frequency-modulation receivers, this chapter would be incomplete without some reference to the manner in which frequency modulation is accomplished. It is not intended to go deeply into the practice or theory of f.m. transmitters, but knowledge of the general principles involved will help to classify the function of certain test instruments to be explained later.

Table 1.1: National v.h.f./f.m. stations of the BBC

Station	Frequencies (MHz)			*E.R.P. each programme
	Radio 2	Radio 3	Radio 4	
Ashkirk, S.E. Scotland	89.1	91·3	93·5 S	18 kW*
Ballachulish	88·1	90·3	92·5 S	15 W*
Ballycastle	89·0	91·2	93·4 NI	40 W*
Barnstaple	88·5	90·7	92·9 W	150 W*
Bath	88·8	91·0	93·2 W	35 W*
Belmont	88·8	90·9	93·1 N	8 kW*
Betws-y-Coed	88·2	90·4	92·6 WH	10 W*
Blaen-plwyf, W. Wales	88·7	90·9	93·1 WH	60 kW
Brecon	88·9	91·1	93·3 WH	10 W*
Bressay, Shetland	88·3	90·5	92·7 S	10 kW*
Brighton	90·1	92·3	94·5 W	150 W*
Brougher Mountain, Enniskillen	88·9	91·1	93·3 NI	2·5 kW*
Cambridge	88·9	91·1	93·3 M	20 W*
Campbeltown	88·2	90·4	92·6 S	35 W*
Carmarthen	88·5	90·7	92·9 WH	10 W*
Churchdown Hill	89·0	91·2	93·4 M	25 W*
Divis, Belfast	90·1	92·3	94·5 NI	60 kW
Dolgellau	90·1	92·3	94·5 WH	15 W*
Douglas, Isle of Man	88·4	90·6	92·8 N	6 kW*
Ffestiniog	88·1	90·3	92·5 WH	50 W*
Forfar	88·3	90·5	92·7 S	10 kW*
Fort William	89·3	91·5	93·7 S	1·5 kW
Grantown	89·8	92·0	94·2 S	350 W*
Haverfordwest	89·3	91·5	93·7 WH	10 kW*
Hereford	89·7	91·9	94·1 M	25 W*
Holme Moss	89·3	91·5	93·7 N	120 kW
Isles of Scilly	89·8	91·0	93·2	20 W*
Kendal	88·7	90·9	93·1 N	25 W*
Kilkeel	88·8	91·0	93·2 NI	25 W*
Kingussie	89·1	91·3	93·5 S	35 W*
Kinlochleven	89·7	91·9	94·1 S	2 W
Kirk o' Shotts	89·9	92·1	94·3 S	120 kW
Larne, Co. Antrim	89·1	91·3	93·5 NI	15 W*
Les Platons, Channel Isles	91·1	94·75	97·1 W	1·5 kW*
Llanddona	89·6	91·8	94·0 WH	12 kW*
Llandrindod Wells	89·1	91·3	93·5 WH	1·5 kW
Llangollen	88·85	91·05	93·25 WH	10 kW*
Llanidloes	88·1	90·3	92·5 WH	5 W
Lochgilphead	88·3	90·5	92·7 S	10 W*
Londonderry	88·3	90·55	92·7 NI	13 kW*
Machynlleth	89·4	91·6	93·8 WH	60 W*
Maddybenny More	88·7	90·9	93·1 NI	30 W*
Meldrum	88·7	90·9	93·1 S	60 kW
Melvaig	89·1	91·3	93·5 S	22 kW*
Morecambe Bay	90·0	92·2	94·4 N	4 kW*
Newry, Co. Down	88·6	90·8	93·0 NI	30 W
Northampton	88·9	91·1	93·3 M	60 W*
North Hessary Tor	88·1	90·3	92·5 W	60 kW
Oban	88·9	91·1	93·3 S	1·5 kW
Okehampton	88·7	90·9	93·1 W	15 W*
Orkney	89·3	91·5	93·7 S	20 kW*
Oxford	89·5	91·7	93·9 M 95·85 W	22 kW*
Penifiler, Skye	89·5	91·7	93·9 S	6 W*
Perth	89·3	91·5	93·7 S	15 W*
Peterborough	90·1	92·3	94·5 M	20 kW*
Pitlochry	89·2	91·4	93·6 S	200 W*
Pontop Pike	88·5	90·7	92·9 N	60 kW
Redruth	89·7	91·9	94·1 W	9 kW*
Rosemarkie	89·6	91·8	94·0 S	12 kW*
Rowridge, I. of W.	88·5	90·7	92·9 W	60 kW
Sandale	88·1	90·3	94·7 N 92·5 S	120 kW
Scarborough	89·9	92·1	94·3 N	25 W*
Sheffield	89·9	92·1	94·3 N	60 W
Skriaig, Skye	88·5	90·7	92·9 S	10 kW*
Sutton Coldfield	88·3	90·5	92·7 M	120 kW
Swaledale	—	—	—	—
Swingate	90·0	92·4	94·4 L	7 kW*
Tacolneston	89·7	91·9	94·1 M	120 kW
Thrumster	90·1	92·3	94·5 S	10 kW*
Toward	88·5	90·7	92·9 S	250 W*
Ventnor, I. of W.	89·4	91·6	93·8 W	20 W*
Weardale	89·7	91·9	94·1 N	100 W*
Wensleydale	88·3	90·5	92·7 N	25 W*
Wenvoe	89·95	96·8	94·3 WH 92·125 W	120 kW
Whitby	89·6	91·8	94·0 N	40 W*
Wrotham	89·1	91·3	93·5 L	120 kW

* Directional aerial system.

Note: At the time of writing some programmes from Wrotham, Brighton, Swingate, Sutton Coldfield, Holme Moss and some of the corresponding relay stations are stereo-encoded.

Table 1.2: Local v.h.f./f.m. stations of the BBC

Station	*Frequency (MHz)*	*Rated E.R.P.*
Brighton ..	88·1	75 W*
Durham ..	96·8	2·6 kW*
Leeds ..	94·6	140 W
Leicester ..	95·05	140 W
Merseyside ..	95·85	2·5 kW*
Nottingham ..	94·8	140 W
Sheffield ..	88·6	30 W*
Rotherham ..	95·05	9 W
Stoke-on-Trent ..	94·9	2·5 kW*

* Directional aerial system.

The simplest way of producing an f.m. signal is by turning the tuning control of an unmodulated signal generator rapidly to and fro. The range over which the tuning is turned corresponds to the deviation, while the speed at which the tuning control is turned corresponds to the modulation frequency. Although this method of producing a frequency-modulated signal may be considered purely illustrative, it may be surprising to know that it was used at one time as the basis of a frequency modulator. Instead of turning the oscillator tuning capacitor by hand, however, a small electric motor was used to spin the variable plates of an auxiliary capacitor which was connected in parallel with the normal tuning capacitor of the oscillator. The auxiliary capacitor was designed so that the oscillator frequency changed linearly with rotation.

The device could not be used for the transmission of audio intelligence but it served the purpose of the more modern frequency-modulated oscillator (usually called a wobbulator) which is used in conjunction with an oscilloscope to provide displays of frequency-response curves of tuned circuits.

Such instruments will be dealt with later, but for the time being it is interesting to note that to secure a display of this nature the horizontal movement of the spot on the oscilloscope's cathode-ray tube (c.r.t.) must be synchronized with the variation in frequency of the oscillator. This is achieved easily enough with the modern wobbulator by electronic means, but with the motor-driven modulator a time-constant circuit (resistor and capacitor) was used which was discharged at the appropriate time by a make-and-break contact operated by the motor shaft.

The simplest possible form of f.m. system for the transmission of audio intelligence comprises a valve oscillator whose frequency is arranged to fluctuate in sympathy with sound waves striking a capacitor microphone connected across the oscillator's tuned circuit. The sound waves cause a change of capacitance proportional to the intensity of the sound, while the rate of change of capacitance is governed by the frequency or pitch of the sound. Since these capacitance changes are reflected across the tuned circuit of the oscillator, the oscillator frequency fluctuates accordingly and its output is frequency-modulated by the sound waves (Fig. 1.6).

This is not a practical arrangement by any means, but serves to provide the initial idea explaining how a frequency-modulated signal is produced. Incidentally, certain broadcast receivers suffer from a defect resulting from frequency modulation of the local oscillator when tuned to a powerful modulated carrier-wave. What happens is that sound waves from the loudspeaker cause the plates

of the tuning capacitor to vibrate, thereby altering the tuning of the receiver in sympathy. This disturbance in turn affects the intermediate-frequency stages and the detector circuit and is reproduced in the loudspeaker. An acoustic feedback path is thus set up between the tuning capacitor and the loudspeaker, which often results in a loud howl.

In ideal conditions, when a set which is susceptible to this effect is tuned to a powerful unmodulated carrier, it is often possible to use the tuning capacitor

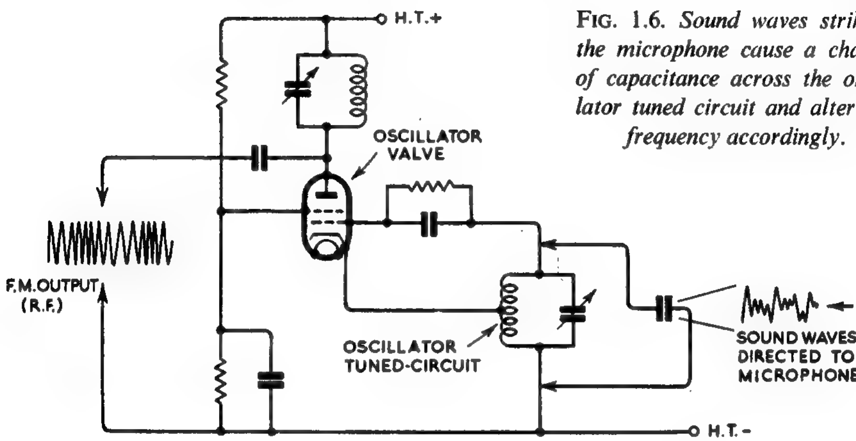


FIG. 1.6. *Sound waves striking the microphone cause a change of capacitance across the oscillator tuned circuit and alter the frequency accordingly.*

as a microphone and to hear the result from the loudspeaker. The effect is known as acoustical feedback, and when present usually indicates that the rubber buffers securing the capacitor to the chassis need replacing. The effect is more noticeable on sets of high selectivity, particularly when tuned to a short-wave station.

REACTANCE MODULATOR

In order to obtain a practical arrangement for frequency-modulating a carrier wave at audio frequencies it is essential for the audio and microphone circuits to be completely isolated and buffered from the r.f. oscillator. An ingenious circuit network which serves this purpose is known as a reactance modulator. It consists, primarily, of a valve, capacitor and resistor which appears to the tuned circuit of the r.f. oscillator as a variable capacitance whose value can be modified by alteration of the valve's mutual conductance (g_m). It will be recalled that the valve's g_m can be altered easily by adjusting the grid-bias voltage. Clearly then, variations of grid-bias will cause corresponding variations of reactance across the tuned circuit of the oscillator and give rise to frequency variations about the nominal frequency of the oscillator. The oscillator frequency will thus, within limits, come under the control of the signal or voltage applied to the grid of the reactance valve.

A simplified reactance-valve circuit is shown in Fig. 1.7. Here, capacitor C1 and inductor L form the essential elements of the tuned circuit of the r.f. oscillator. The parallel-connected valve, in conjunction with a resistor-capacitor combination R and C2, appears to the oscillator tuned circuit as a capacitance;

that is, it presents a 90-deg. phase shift in the capacitive sense—the current flowing through the valve leads the voltage across it by a quarter of a cycle (90 deg.).

If the action of the valve is first considered without the associated R and C2, we can clearly visualize the oscillator voltage being in phase with the resulting

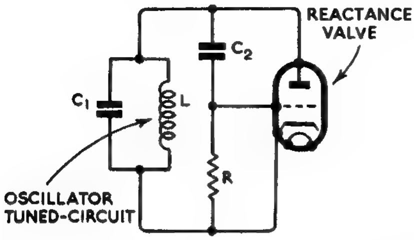


FIG. 1.7. *Reactance-valve circuit.*

current through the valve—assuming that the valve is in receipt of a high-tension (h.t.) and a low-tension (l.t.) supply, but not shown in Fig. 1.7. The valve on its own, therefore, appears to the oscillator circuit as a pure resistance, the current and voltage being in phase.

Now, if the added effect of R and C2 be considered, we shall observe that a part of the oscillator voltage is applied to the grid of the valve, but its phase is advanced by nearly 90 deg. due to the effect of the capacitor. The voltage is amplified by the valve and produces in the anode circuit an oscillatory current component leading by 90 deg. on the applied oscillator voltage which, of course, is equivalent to a capacitive shunt across the oscillator circuit. The magnitude of capacitive reactance so produced is readily altered by modifying the effective g_m of the valve, for example, by varying the control-grid bias potential.

This somewhat practical explanation (a precise analysis of the circuit cannot be given without recourse to mathematics) is representative of the action of nearly all types of electronic reactance circuits, there being several deviations from the simplified example. It is interesting to note, however, that by transposing the capacitor and resistor the circuit is made predominantly inductive—an arrangement sometimes utilized.

When used for the production of a frequency-modulated signal, the g_m of the reactance valve is arranged to fluctuate at a rate equal to the frequency of the modulation signal and at a magnitude proportional to the intensity of the modulation. The circuit thus serves to convert a change in audio-frequency voltage into a change of capacitance or inductance.

The circuit in Fig. 1.8 depicts a typical reactance-modulator circuit coupled to an r.f. oscillator. Here will be seen the reactance valve in parallel with the oscillator tuned circuit, via the coupling and isolating capacitor C2. Capacitor C1 and the series-connected resistors R1 and R2 form the reactance-valve components. The a.f. modulation signal is applied to the junction of the resistors R1 and R2, thereby causing the apparent capacitive reactance across the oscillator circuit to fluctuate in sympathy with the modulation signal; thus, frequency modulation is achieved.

Although the general principle remains the same, in practice the circuitry associated with f.m. transmitters is considerably more complex. In order to

maintain frequency stability of the master oscillator, a sample of the fully modulated frequency-modulated signal is taken, and the nominal frequency compared with a signal of fixed and stabilized frequency derived from a crystal oscillator. A discriminator circuit is adopted to compare the two signals and to give zero output when their frequencies coincide.

When a difference in frequency occurs, however, due to a drift of the master-oscillator, a positive or negative bias is produced by the discriminator. This is used as a control potential for a reactance-valve circuit which is connected across the tuned circuit of the master oscillator. Thus, a frequency drift of the master oscillator will produce a d.c. bias of correct polarity to swing the virtual inductive or capacitive reactance of the reactance valve in the direction required to bring the frequency of the master oscillator in line with that of the crystal oscillator. This is known as the *centre frequency stabilized system*. There are many other specialized circuit arrangements evolved essentially for f.m. transmitters, but there is little need to deal with them here.

Swings in frequency can also be given by ferrite modulators and capacitor-diodes, the latter being considered in Chapter 4.

It is of interest to note that the master oscillator of the transmitter usually functions at a sub-multiple of the ultimate nominal carrier frequency. A number

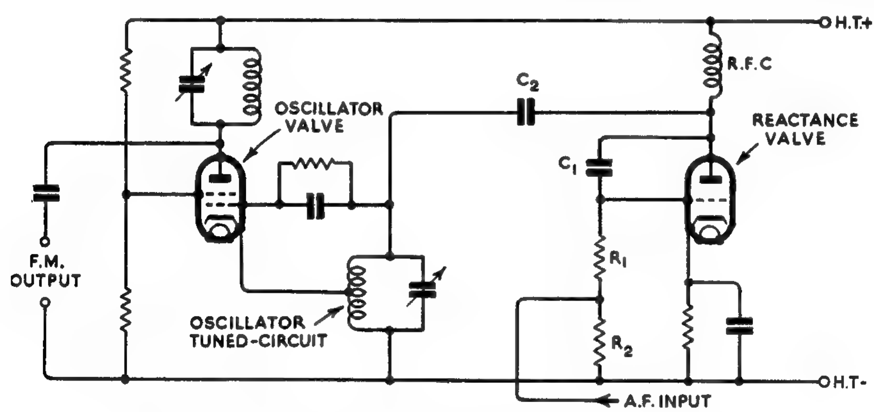


FIG. 1.8. *Reactance modulator circuit coupled to an r.f. oscillator.*

of frequency-multiplier stages are incorporated which step up the relatively low frequency of the master oscillator. In this connexion it will be realized that such stages will also multiply the deviation frequency, thereby reducing the deviation required by the reactance-modulator circuit.

POLARIZATION OF F.M. SIGNALS

The characteristics of the f.m. signal and its method of production have now been studied so that it is appropriate to have some idea how the signal is radiated from the transmitting aerial. The v.h.f./f.m. radio transmitters of the BBC use horizontal polarization.

Radio waves consist of two components, an electric component (electric field) and a magnetic component (magnetic field); these components are alter-

nating at the same time in the direction of wave travel. They are not alternating in the same plane, though, but at right-angles to each other.

The character of radio waves is often depicted in similitude to the waves which spread outwards when a pond of still water is disturbed by a stone, or as the waves which occur on a long outstretched rope which is vigorously waved either up and down or from side to side. These are good analogies within limits, particularly the first one where the disturbance to a floating object, such as a cork, is compared with the effect radio waves have on a receiving aerial. The cork bobs up and down as each wave passes under it, but does not travel across the surface with the wave. Extending the analogy, the powerful signal conveyed to the transmitting aerial causes an electromagnetic disturbance much in the same way as the stone disturbs the water, and radio waves are set up which travel outwards in all directions from the aerial. Receiving aerials in range of the waves are thus excited and weak signal voltages are set up in them which are passed on to the receiver.

Neither of these analogies gives a true picture concerning the polarization of the wave. The diagram in Fig. 1.9 gives a more tangible representation—if such

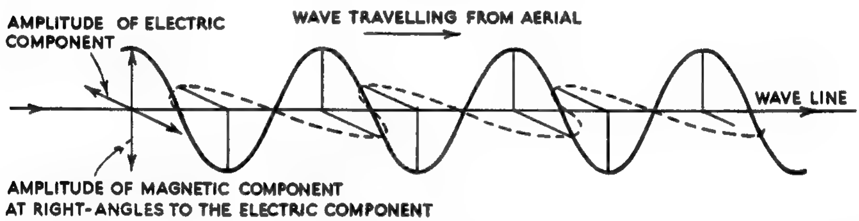


FIG. 1.9. *The electric and magnetic components of a radio wave. The wave is polarized in the direction of the amplitude of the electric component, which is in the horizontal sense in this diagram.*

can be said of an invisible force—of the happenings in space on one line away from the aerial. Here is illustrated the alternations of the electric and the magnetic components of the wave which are at right-angles to each other.

Now, the important thing here is that the signal is polarized in the direction of the amplitude (alternations) of the electric component, which, so far as Fig. 1.9 is concerned, is in the horizontal sense. With an ordinary dipole aerial (to be discussed later) connected to a transmitter, the radiated waves are vertically polarized when the aerial is positioned vertically, and to pick up such waves the receiving aerial must be positioned vertically. Conversely, a horizontally positioned transmitting dipole aerial will send out horizontally polarized waves which require a horizontally positioned receiving aerial for optimum response to the radiated waves.

There is no need to go into details concerning how this polarized disturbance in space occurs, for the present purpose it is sufficient to know that a vertical dipole responds to vertically polarized signals and a horizontal one responds to horizontally polarized signals.

As radio waves travel through space the polarization can sometimes alter from that intended at the transmitting aerial. This effect is caused by atmo-

spheric disturbances and bending of the waves as they travel. It is for this reason that BBC frequency-modulated signals are sometimes picked up quite well on a television aerial which is vertically polarized.

There is a type of aerial, known as the slot aerial (to be discussed later), which polarizes the wave in the opposite sense to the conventional dipole aerial. With this aerial a slot of suitable size cut in a piece of wire netting functions as the dipole, and needs to be vertically positioned to radiate and receive a horizontally polarized signal.

COMPARISON OF RECEIVING SYSTEMS

For the reception of both a.m. and f.m. signals, the signals picked up by the aerial are taken to the first stage of the receiver (the r.f. amplifier) for initial amplification at signal frequency.

Owing to the relatively low frequencies of the long-wave and medium-wave broadcast bands, a simple end-fed wire aerial, or even an odd piece of wire, is usually all that is required to pick up a sufficient signal for application to the r.f. amplifier. At the higher frequencies of Band II, however, an aerial of more critical proportions is demanded. As previously intimated, a dipole aerial of some sort is needed. Whereas an ordinary end-fed wire aerial, as used on the l.w. and m.w. broadcast bands, has no definite response at any particular frequency in these bands, at least not intentionally, the dipole is a tuned aerial and has a response endowing it with certain characteristics (to be considered later) within the frequency spectrum of 87 to 108 MHz comprising Band II.

With f.m. receivers, therefore, the aerial serves as a tuned circuit as well as a collector of signals. A simple piece of wire can be used with f.m. receivers in areas where the signal is strong, but for best results in average or low-signal-strength areas, the extra magnification given by the dipole is nearly always called for.

In both receiving systems the amplified signal at the output of the r.f. stage is applied to the frequency-changer stage the function of which is to change the amplifier signal frequency to a lower frequency so that it can be more easily handled and amplified. This lower frequency, in both systems, is known as the intermediate frequency (i.f.).

One of the essential differences between the two receiving systems lies in the r.f. and i.f. amplifier stages. It will be recalled that an amplitude-modulated signal carries components of the modulation signal, known as sidebands, which alternate above and below the nominal frequency of the carrier wave (Fig. 1.3). These sidebands due to a modulation signal of, say, 10 kHz in relation to a carrier frequency of 1,000 kHz thus extend from 990 kHe to 1,010 kHz.

Now, in order to maintain the correct amplitude of the 10 kHz modulation signal in relation to other modulation frequencies comprising a complex sound waveform, the r.f. and i.f. amplifier stages must respond equally to the sideband components up to the sideband produced by the highest audio-frequency it is required to receive.

Fig. 1.10 (a) depicts a typical response curve relating to an r.f. and i.f. stage of an a.m. receiver. Here it will be seen that a signal applied at carrier frequency (1,000 kHz) gives rise to maximum output from the stages, while the output due

to the sideband components of the signal (still assuming 10 kHz modulation) is a little more than half that at carrier frequency. The output due to sideband components created by modulation frequencies below 10 kHz is correspondingly larger, while the output due to sideband components above 10 kHz is correspondingly smaller.

It is seen then, that the sideband components of the signal become progressively attenuated as the modulation frequency is raised. Fig. 1.10 (b) shows an

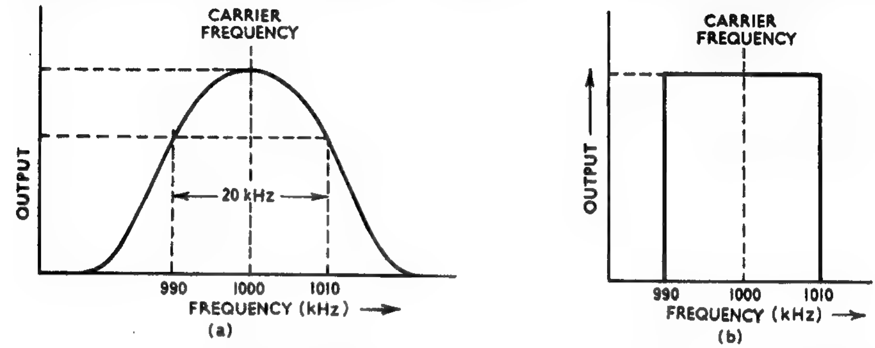


FIG. 1.10. *(a) Typical response surve of an a.m. receiver; (b) ideal form of response curve which would maintain equal amplification at all modulation frequencies up to 10 kHz and then cut off completely.*

ideal form of response curve which would maintain equal amplification at all modulation frequencies up to 10 kHz and then cut off completely. Such a response is virtually impossible to obtain in practice, so that it is necessary to compromise and secure a response curve most suitable for the purpose in mind.

Owing to the close spacing of stations in the m.w. and l.w. broadcast bands, the response has to be purposely narrowed to prevent adjacent stations from interfering with the one to which the receiver is tuned. This narrowing of the response thus limits the quality of audio reproduction by severely attenuating the higher audio-frequency components of the modulation signal. At the present time, owing to serious congestion on the m.w. and l.w. broadcast bands, one is fortunate if able to reproduce at the correct relative level any audio frequency above about 5 kHz without whistles and sideband-splash from adjacent stations marring the reception. Indeed, even if an a.m. receiver possesses a fairly wide pass-band, it is usually found necessary to cut the higher audio frequencies, as a means of suppressing the interfering whistles, by turning the tone control.

With f.m. receivers, the wider spread of the sidebands and the relatively large deviation frequency for 100-per-cent modulation calls for a receiver r.f. intermediate-frequency pass-band considerably wider than is required for the a.m. counterparts. Roughly, the *total* pass-band required by an f.m. receiver is equal to twice the peak deviation plus twice the modulation frequency. Thus, for 100-per-cent modulation (plus and minus 75 kHz) at a modulation frequency of 10 kHz the required total pass-band becomes 170 kHz. A response curve satisfying this condition is shown in Fig. 1.11. In practice, quite a reasonable output would be obtained at 100 kHz above and below the carrier frequency, so the average pass-band of such a response would be assessed at about 200 kHz,

which is ten times that of the a.m. case for the same modulation frequency!

The design of the f.m. intermediate-frequency tuned circuits, coupled with the fact that the intermediate-frequency is higher with f.m. receivers than with

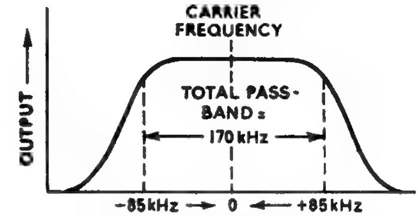


FIG. 1.11. *A total pass-band in the region of 170 kHz is necessary to carry an f.m. signal modulated at 10 kHz.*

a.m. ones, readily permits the formation of the relatively wide f.m. overall response curve. It will also be recalled that the f.m. channels are about 200 kHz wide so that, even if the overall response spreads out over 170 kHz, which is usually arranged anyway as a means of nullifying the effect of local oscillator drift in the receiver, there is very little fear of adjacent channel interference resulting as would be the case in the a.m. channels where spacing between stations is sometimes less than 9 kHz.

It will be seen later that the responses of individual stages are adjusted so that together they merge to form the ideal overall response which is essential for high-quality reception.

DEMODULATION OR DETECTION

Undoubtedly, the chief difference between the two receiving systems lies in the detector stage. Provided that the r.f. and i.f. amplifying stages of a receiver are sufficiently broadly tuned to pass all the sidebands of significant strength, an f.m. signal will be handled by them in an identical manner to an a.m. one. On arrival at the detector stage, however, the f.m. signal receives somewhat specialized treatment.

In an a.m. receiver the amplitude-modulated carrier wave, either at signal frequency or at intermediate frequency, is applied to a simple half-wave rectifier. The modulated carrier wave is thus rectified, giving rise to a current in the

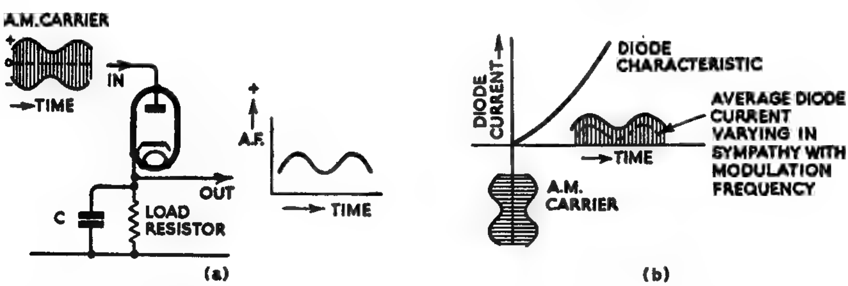


FIG. 1.12. *(a) Simple diode a.m. detector circuit; (b) operating characteristics of the diode.*

associated load resistor; this current develops across the load a d.c. voltage on which is superimposed an audio-frequency voltage corresponding to the modulation signal. This is illustrated in Fig. 1.12; at (a) is shown the circuit, in which

capacitor C eliminates the r.f. component of the signal after rectification; at (b) the function of the diode is depicted.

It is clear from these illustrations that if an f.m. signal is applied to an a.m. detector of this kind, all that will be obtained across the load resistor is a steady d.c. voltage corresponding to the average value of the rectified carrier signal; the amplitude of the signal remains constant with modulation, so no audio-frequency voltage will occur across the load resistor.

There is an artifice which can be used to permit the reception of an f.m. signal of an a.m. receiver. Let us look again at a response curve of a receiver (Fig. 1.13), and let it be supposed that the receiver is tuned exactly at, say, 10 MHz, as indicated, and that an applied f.m. signal has a deviation of plus and minus 100 kHz. The signal is thus varying from 9·9 to 10·1 MHz and the output from the tuned stages of the receiver will increase from 9·9 to 10 MHz and decrease from 10 to 10·1 MHz. If the rate of deviation corresponds to an audio-frequency signal, and the receiver has a simple a.m. detector, then an a.f. signal will appear across the detector load resistor. What has happened is that the f.m. signal has been changed to an a.m. one and then demodulated by the a.m. detector in the normal manner.

This system has no practical value, though, as revealed by the diagram (Fig. 1.13), the modulation evelope given to the original f.m. signal now corresponds

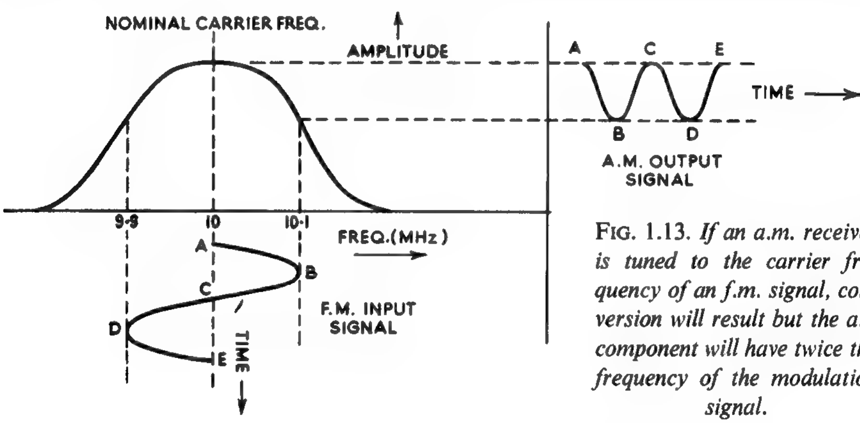


FIG. 1.13. *If an a.m. receiver is tuned to the carrier frequency of an f.m. signal, conversion will result but the a.f. component will have twice the frequency of the modulation signal.*

to twice the modulation frequency. This is clearly realized by following the points marked on the input and output waveforms.

If an a.m. receiver is slightly detuned so that the nominal frequency of the f.m. carrier falls on the sloping side of the response curve, then the modulation frequency is unaffected and simple f.m. detection is achieved. This process is illustrated in Fig. 1.14, from which will be seen that a change in frequency of plus and minus 100 kHz gives rise to a corresponding variation in amplitude, as governed by the selectivity characteristics of the receiver's tuned circuits. The frequency swing of the carrier as it is modulated moves the carrier up and down the slope of the response curve, thereby producing a greater or less instantaneous voltage across the output of the r.f./i.f. amplifying stages as the signal approaches or recedes from the tuned frequency.

The receiver used for this purpose would undoubtedly be of the superhet mode, so the resulting amplitude fluctuations would be in i.f. voltage, which would arrive at the a.m. detector and be treated in exactly the same way as an ordinary a.m. signal. An audio-frequency output corresponding to the modulation signal would thus appear across the detector load resistor.

The success of this system depends to a large degree on the selectivity characteristics of the tuned circuits for clearly, if the sloping side of the response

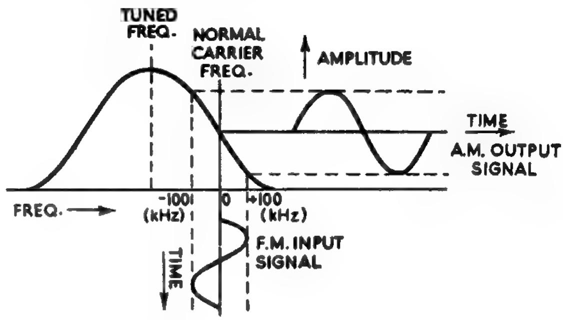


FIG. 1.14. *By slightly detuning the receiver so that the nominal carrier frequency of the f.m. signal falls on the sloping side of the response curve, f.m. signals are converted to a.m. signals, the modulation frequency being unaffected.*

curve is far from linear then considerable distortion will result. Moreover, amplitude disturbances of the carrier would be accepted by the receiver and extreme difficulty would be encountered in tuning (or detuning) the receiver accurately so that the nominal carrier frequency falls exactly in the centre of the most linear portion of the sloping side of the response curve.

For these reasons this system cannot be used for quality reception.

THE A.F. SECTION

Obviously, f.m. detectors of somewhat more elaborate design are employed in sets of modern styling, and these will be completely covered in a later section. However, it is the aim in this chapter to obtain a general idea of the f.m. signal, how it compares with the a.m. signal, and of the factors which are common between the two receiving systems. By pursuing this course it will be found easier to assimilate the functions and characteristics appertaining to individual f.m. sections, when breaking away from the a.m. parallel which so far has been followed.

After leaving the detector, the audio-frequency of the f.m. signal, as with the a.f. component of the a.m. signal, is applied to an a.f. amplifier to increase its amplitude and give it characteristics of power so that it is suitable for operating the loudspeaker.

Owing to the extended frequency range which can be carried by an f.m. system, it is usually found that the a.f. stages employed in f.m. receivers possess a wider response to audio frequencies than the equivalent stages in receivers employing a.m. only. It would be pointless to feature an a.f. section in an a.m. receiver which section faithfully amplifies all frequencies from the lowest up to some 10 to 15 kHz when the pass-band of the i.f. section is limited to 5 kHz or so.

With f.m., things are different as already seen, so with good purpose the a.f. section can be designed to embrace audio frequencies up to at least 15 kHz.

The response of the a.f. section to the higher audio-frequency signals permits the reproduction of harmonic components of the original sound, and it is the reproduction of these which makes for high-quality reception.

As a means of assisting in this respect, two or more loudspeakers are sometimes used in f.m. sets. The main loudspeaker is employed for the reproduction of sound corresponding to the low and middle registers, while a very small loudspeaker serves to reproduce the very high audio signals which are not normally heard on receivers employing a.m. only.

Where two loudspeakers are used, the one handling the low and middle audio signals is called the bass unit, while the smaller one for the higher frequencies is called the treble unit.

There are two additional features of the f.m. system which have no a.m. parallel. One is that the f.m. modulation signal is purposely 'distorted' at the transmitter in such a way to produce a rising-frequency characteristic resulting in greater amplification of the higher modulation frequencies in relation to the lower ones.

At the receiver, circuits are employed to correct for this 'distortion' after the signal has been demodulated. This may seem all rather pointless, but it will be seen in the next chapter why it is done.

The second feature found in f.m. receivers but not in a.m. ones is the amplitude limiter. As we are not interested in amplitude fluctuations of the carrier wave before it is applied to the detector stage, it is often found advantageous to make use of a simple circuit arrangement whose purpose is to shave off any amplitude disturbances which may occur on the carrier wave as it passes through the various stages of the receiver, and which may have been picked up by the aerial together with the f.m. signal.

CHAPTER 2

The Advantages of F.M.

WITH a good-quality commercial receiver or a tuner unit coupled to an audio amplifier, one has only to tune from say, Radio 2 on the l.w. broadcast band to Radio 2 on the v.h.f./f.m. band to appreciate the vast improvement in reproduction which is given by the f.m. system. With combined a.m./f.m. receivers such a comparison is easily made, and the full force of the difference in quality will be realized if the programme—which is radiated simultaneously on both systems—is "live". The comparison may not be so striking if the programme is recorded, owing to the frequency limitations of the recording itself and of the ancillary playback equipment, or if station links and landlines restrict the a.f. range.

Musical instruments which appear to be "muddled" when heard on the a.m. system are brought clearly into "focus" when the changeover to f.m. is made. This applies particularly to instruments which primarily occupy the upper part of the audio-frequency spectrum. The simple triangle, for example, might well be non-existent when listening on the a.m. system, but can usually be heard with remarkable fidelity when the same passage is received on the f.m. system. Transient instruments, such as the cymbal, since their true tone is given by harmonics extending to the upper limit of the audio spectrum, are very clearly reproduced; in fact, the whole orchestral rendering appears to take on a new brilliance which cannot be equalled on the m.w. and l.w. amplitude-modulated system.

So-called tone-compensating circuits and tone controls are employed in some a.m. receivers to reduce the effect of interfering whistles and crackles, and not to enhance the quality of reproduction. They are purely palliatives which have, unfortunately, become accepted as part of the radio set by a large number of listeners. As will be substantiated by quality-conscious service engineers on returning a radio set after repair, it is often intimated by the owner that the set does not sound the same as hitherto—that the "depth of quality" is missing! The knowledgeable engineer, realizing the trouble, delights his customer by turning on the tone control and cutting the a.f. response above about 2 kHz!

Although the general listener is now becoming quality conscious, probably aided by the improved a.f. response of the TV sound channel, there are still many listeners to whom a good top response is considerably less important than plenty of bass. With this listener, the present-day radio salesman finds it extre-

mely difficult to demonstrate convincingly the much improved quality of the f.m. system, particularly if, as is usually the case, the receiver about to be replaced has been used by the listener for many years.

The f.m. system in itself does not provide for enhanced quality of reproduction over the a.m. system. The ultimate sound that is given by the loudspeaker is considerably coloured by the transmitting system, the transference of the signal from the transmitting aerial to the receiving aerial and by the passage of the signal through the receiver. Many factors are involved, and it is their cumulative effect which governs the quality of the reproduced sound.

The extended bandwidth of the f.m. system readily permits the conveyance of all audio signals necessary for high-quality broadcasting. Nevertheless, such would be the case with a wide-band a.m. system given the same interference-free channels to operate in. There is, though, the question of interference picked up with the required signal at the aerial. With the a.m. system such disturbances are not easily eliminated as they tend to become superimposed on the envelope of the carrier wave in the form of sharp spikes (Fig. 2.1). These spikes of interference look to an a.m. receiver very much like ordinary amplitude modulation

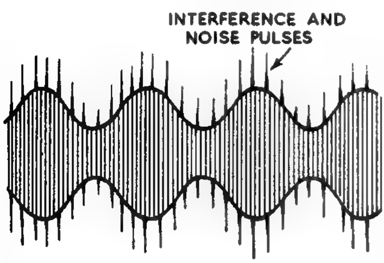


FIG. 2.1. *Impulsive interference and receiver noise appear as sharp spikes on the envelope of the carrier wave.*

and are treated as such by the detector stage and they give rise to staccato bursts of noise from the loudspeaker.

Domestic electrical appliances such as hair dryers, vacuum cleaners and shavers give rise to this kind of interference on the l.w. and m.w. broadcast bands as well as on the v.h.f. frequency-modulated band, so the use of a.m. in Band II would not solve the problem of interference. Indeed, in the v.h.f. bands interference from car-ignition systems also becomes troublesome, though it is not very often heard in the l.w. and m.w. bands. In this respect, therefore, the f.m. system is advantageous because the receiver is not designed to respond to amplitude variations of the carrier.

There is also the question of circuit and valve or transistor noise which is actually generated in the receiver itself. Again, this disturbance tends to provoke amplitude fluctuations of the signal as it passes through the various amplifying stages of the receiver and is invariably heard in the form of a hiss, much like escaping gas. If the signal at the aerial is weak, then the receiver noise outweighs the wanted signal and gives that disturbing background effect to the reproduction. The effect is aggravated as the operating frequency is increased, so will be more troublesome on Band II than on the l.w. and m.w. bands, especially on stereo with a weak aerial signal, but it should be noted that low-noise bipolar and f.e.t. devices at the front-end instead of valves considerably improve the

noise performance. The use of an a.m. system on Band II, then, would not help from this aspect either.

Whistles produced on an a.m. system are usually caused by an unwanted signal heterodyning or beating with the wanted signal and producing amplitude fluctuations at a rate equal to the difference between the frequencies of the two signals. If this difference frequency falls within the audio spectrum and within the pass-band of the receiver, then it is reproduced as a whistle. This is the beat note of the two frequencies; the f.m. system helps to eliminate this as the receiver is not designed to respond to a.m. fluctuations.

FACTORS RELATING TO INTERFERENCE

Although an f.m. receiver is not designed to give an output to a.m. fluctuations of the applied signal, the amplitude disturbances unfortunately give rise to a so-called "by-product" f.m. component on the signal as it makes its way to the detector stage.

Before this effect can be discussed further it will be necessary to get a rough idea regarding phase relationships between two signals or waveforms. In Fig. 2.2 (a), let the full-line waveform represent a signal of given frequency and amplitude. Relatively, then, the broken-line waveform can be seen to be of smaller amplitude, though of the same frequency. Moreover, the peaks of the

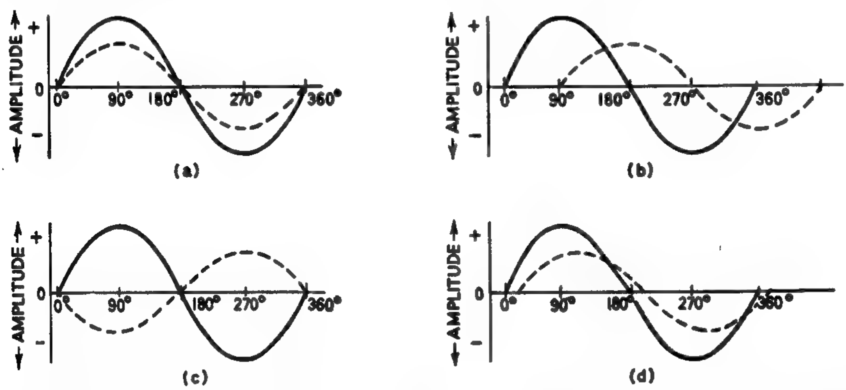


FIG. 2.2. *Phase relationships between two signals: (a) in phase; (b) 90 deg. out of phase; (c) 180 deg. out of phase; (d) a few degrees out of phase.*

broken-line waveform occur at the same time as those of the full-line waveform. This factor indicates that the two signals are in step or in phase. Thus, there are three primary factors, frequency, amplitude and phase.

In Fig. 2.2 (b), a 90 deg. phase difference exists between the two signals. Here it will be seen that one wave is at its peak while the other is at zero. In Fig. 2.2 (c) the phase difference is 180 deg., that is, the positive peak of one wave is occurring at the same instant as the negative peak of the other wave. Phase is also sometimes expressed as a fraction of a cycle instead of in degrees. Using this method, it is correct to say that one wave with respect to the other in Fig. 2.2 (b) is either leading or lagging by a quarter of a cycle. Similarly, a lead or lag of half a cycle is illustrated at (c).

The angular method of expression is usually more convenient to adopt when the phase variation between two waves is small. For example, the relative phase of the two signals in Fig. 2.2 (d) is a matter only of a few degrees, so that it cannot easily be expressed as a simple fraction of a cycle. The angular scale extending from a relative zero to 360 deg. is a measure of phase over one complete cycle of signal; over one-and-a-quarter cycles the measure would be 450 deg. and over one-and-a-half cycles 540 deg., and so on.

The phase difference of two signals can best be appreciated by the use of vector diagrams. Vectors are represented by means of straight lines terminating in arrow-heads. The length of the line indicates the relative magnitude of the quantity it represents, and the direction of the line with its arrow-head shows the direction in which the vector is operating. A vector thus indicates a quantity possessing both magnitude and direction.

In Fig. 2.3 vectors are used to illustrate the phase difference between the two signals or waveforms S1 and S2. The signals are represented by two vectors

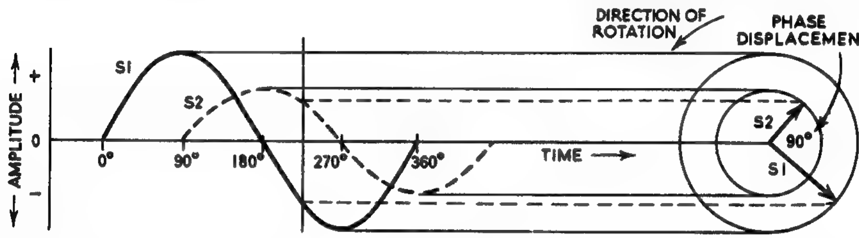


FIG. 2.3. *Here, the phase displacement between signals S1 and S2 is shown by the vectors S1 and S2.*

which can be considered as pivoted on the datum line within the two circles whose diameters are representative of the peak-to-peak amplitude of the signals.

It should be supposed that the two vectors are rotating in an anti-clockwise direction at a constant speed such that one complete revolution—360 deg.—represents one complete cycle of signal. A study of this diagram will make it clear that the angle between the vectors corresponds to the phase displacement between the signals. Now, if the signals are equal in frequency, but are different only in phase (and probably amplitude), then the phase displacement between them will remain constant with time—one vector will lead or lag the other at a fixed angle.

If, on the other hand, the signals differ in frequency, it follows that the phase displacement will be continuously altering with time. The vector corresponding

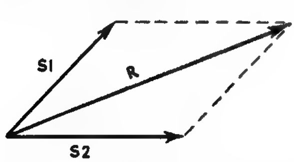


FIG. 2.4. *The resultant of two vectors corresponding to signals S1 and S2 may be found by completing the parallelogram and measuring the resultant R.*

to the higher-frequency signal will rotate at a greater speed than the one corresponding to the lower-frequency signal.

Before this information is put to practical use, however, it will help to understand that the resultant voltage due to the two signals, or due to any two or more

signals which are not in the same phase, can also be computed by means of vectors. The resultant of the two vectors corresponding to signals S1 and S2 in Fig. 2.3 may be found by completing the parallelogram and taking the resultant R, as illustrated in Fig. 2.4.

Another way is by adding the vectors in tandem and taking the resultant from the beginning of one to the end of the other. In the diagram of Fig. 2.5, vectors

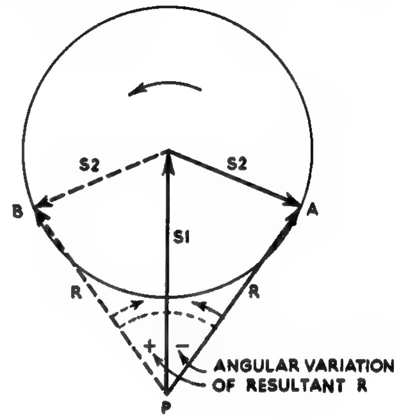


FIG. 2.5. *The resultant voltage of two signals, represented by vectors S1 and S2, may be found by adding the vectors and taking the result from the beginning of one to the end of the other. The resultant component R changes in magnitude and phase as the phase of S1 changes relative to S2.*

S1 and S2 represent two signals which differ in phase, and by adding them in tandem the resultant is shown by the dimension R.

Now, if the phase of one signal relative to the other is changing, as just discussed, for the purpose of illustration it can be considered (Fig. 2.5) that S1 is fixed and that S2 is rotating anti-clockwise. Thus, where the signals differ in frequency the magnitude of the resultant signal R will change from S1 minus S2 to S1 plus S2, as the phase of S1 changes relative to S2.

Apart from a variation of magnitude, the diagram also reveals that the phase of the *resultant* signal oscillates to and fro, relative to S1, as the phase of S1 and S2 changes. The broken lines show the conditions which exist when S2 has made just over half a revolution (about 200 deg.) with respect to S1.

The resultant quantity R can, therefore, be considered as pivoted at point P and oscillating backwards and forwards between points A and B as the phase between the signals changes. The angular variation of R will thus range from a negative value to a positive value depending upon the relative strength of S1 and S2.

Now let us look at this first in relation to an interfering unmodulated r.f. signal whose frequency is slightly removed from the carrier frequency of the wanted signal. As already seen, the interfering signal and the wanted signal produce a resultant signal whose amplitude varies from the sum to the difference of the two vector values at a rate equal to the difference in frequency, thereby giving amplitude modulation. The depth of the modulation so produced is proportional to the relative strengths of the interfering and wanted signals.

Apart from provoking amplitude disturbances, however, it is also seen that the interfering signal affects the phase angle of the resultant signal and causes it to oscillate from a negative to a positive value.

Since a change of phase is virtually the same thing as a change of frequency, and if the phase change occurs regularly, as in the case under consideration when two signals of fixed frequency combine, the overall effect is equivalent to frequency modulation. Because a phase change of the resultant signal over a period of time increases in relation to the frequency difference of the interfering and wanted signals, the amount of frequency modulation so produced also increases as the frequency difference increases.

IMPULSIVE INTERFERENCE

Consider what takes place when the interference is of the *impulsive* kind, such as that produced by car-ignition systems and electric motors. From Fig. 2.1 it has been seen that this kind of interference becomes superimposed on the envelope of the carrier wave in the form of sharp spikes. These spikes or pulses last only for a small fraction of a second, and since they are sharp-edged and of a transient nature they are composed of a very wide spectrum of frequencies extending from very low frequencies up to hundreds of megacycles per second. In this form they are not directly accepted by the receiver, which has a pass-band relatively very much limited.

Nevertheless, on arriving, together with the signal, at the first tuned circuit each pulse causes the circuit to oscillate or "ring" at its natural frequency (that is the frequency to which it is tuned), in much the same way as the string of a harp vibrates at its tuned frequency when plucked. With the tuned circuit, of course, the oscillations are purely electrical as distinct from the mechanical vibrations of the harp string.

A train of damped oscillations is thus developed within the tuned circuit (as shown in Fig. 2.6) and on passing through the r.f./i.f. amplifying stages, the

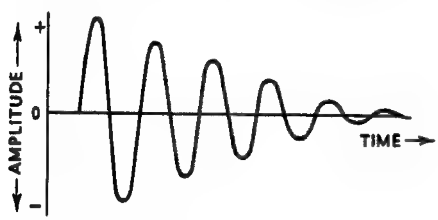


FIG. 2.6. *A train of damped oscillations resulting from an interference pulse. If the receiver is accurately tuned to the required signal, the resultant signal in the receiver will have no f.m. component.*

tuned circuits following are caused to oscillate in sympathy at increased amplitude. With a superhet receiver the damped oscillations will be converted to the intermediate frequency together with the signal.

Essentially, then, impulsive interference also gives rise to conditions more or less equivalent to those which were discussed in relation to an interfering r.f. signal. With impulsive interference, however, the r.f. signal which is created by the interference pulses diminishes in amplitude and is often less troublesome. Moreover, if the receiver is properly aligned and tuned accurately to the wanted signal, the frequency of the damped oscillations coincides with the frequency of the signal. This means that no by-product f.m. is produced, and if the receiver is in good condition little or no output, due to the interference, occurs at the loudspeaker.

Impulsive interference cannot be tuned out because its secondary r.f. effect

takes on the frequency to which the receiver is tuned. It is interesting to note that the duration of the train of damped oscillations is somewhat influenced by the overall pass-band of the receiver. Contrary to expectations, the oscillations last longer in receivers with a restricted pass-band than in relatively wide-band receivers, such as the vision channel of a television set.

This is a potential advantage in favour of the f.m. case, since a relatively wide pass-band is necessary to cater for the large deviation and the extended side-bands. In an a.m. receiver with an overall pass-band of, say, 30 kHz (this is quite wide for a.m. and would necessitate operation in the v.h.f. region) a train of damped oscillations caused by an interference pulse lasts about six times as long as that caused by a similar pulse in an f.m. receiver having an overall pass-band of 200 kHz.

So far, r.f. and impulsive interference has been dealt with. There is still the question of mutual interference caused by stations operating on the same frequency or slightly displaced in frequency. Listeners are well aware of the disturbance which such conditions have on the m.w. and l.w. amplitude-modulated system. Unless the required station is giving an aerial signal in the region of 25 times that of the signal of the interfering station, it cannot truthfully be said that the result is of entertainable value. This effect can readily be demonstrated by tuning in an interference-free station near the high-frequency end of the m.w. band after dark.

THE CAPTURE EFFECT

With f.m. things are not nearly so bad as this. In fact, only a small disturbance is produced by two mono stations working in the *same* channel provided the aerial signal of the wanted station is little more than double the aerial signal of the unwanted station. This is known as the *capture effect* (related to a given signal/noise ratio and expressed in decibels) and represents one of the most desirable features of a wide-band f.m. system. It makes it readily possible for channel sharing without the necessity of taking extreme precautions so far as signal polarization and unsettled atmospheric conditions are concerned, as is the case with television. With f.m., each transmitter of a common-frequency network can be situated to have its own zone providing relatively interference-free reception.

In-channel and adjacent channel stereo stations, however, can override the mono capture ratio to some extent, depending on the strength of the aerial signal (see Chapter 7).

AMPLITUDE LIMITING

Nearly all f.m. receivers have an amplitude-limiter stage whose purpose is to shave off interference pulses and amplitude fluctuations of the f.m. carrier before it is applied to the detector stage. Seeing that the detector is not concerned in amplitude modulation and that any such disturbance of the f.m. signal is liable to provoke unnecessary noise, it is permissible to limit the amplitude of the signal to some pre-determined level irrespective, within limits, of the strength of the aerial signal. The limiting or cutting of the signal in this way does not detract from the quality of the audio intelligence.

The limiting action is ingeniously though simply achieved in the i.f. amplifier prior to the detector stage. A method of doing this in a valve receiver is by operating the valve so that it has a small grid base. This simply means that the valve is pushed into the region of anode-current cut-off when the negative voltage at the grid exceeds one or two volts. This condition is brought about by the artifice of running the valve with an abnormally low screen-grid voltage. Moreover, the control grid is connected to a resistor-capacitor combination after the style of a leaky-grid detector, so that the greater the signal applied to the grid the greater the negative voltage developed across the capacitor and the more the signal is pushed into the anode current cut-off region.

The action of the circuit is shown in Fig. 2.7, from which it will be seen that all amplitude disturbances are removed from the signal and that perfect limiting

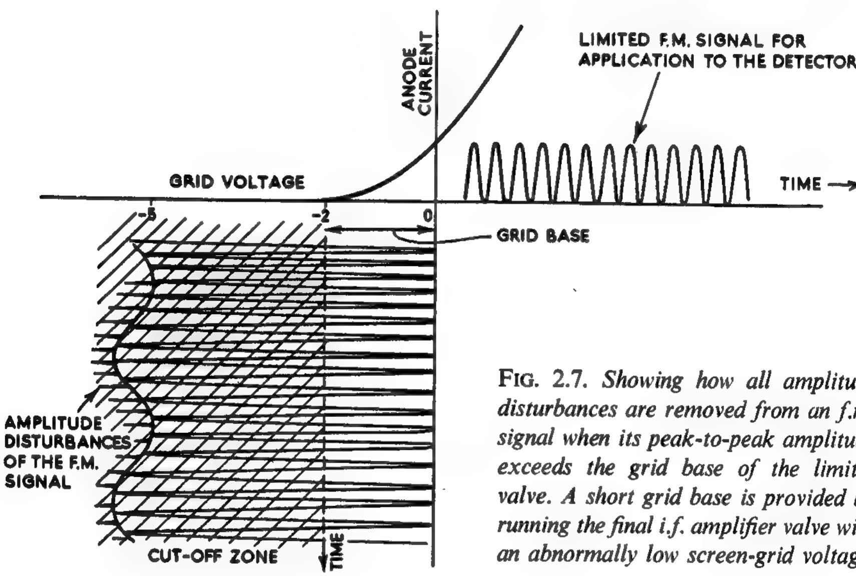


FIG. 2.7. *Showing how all amplitude disturbances are removed from an f.m. signal when its peak-to-peak amplitude exceeds the grid base of the limiter valve. A short grid base is provided by running the final i.f. amplifier valve with an abnormally low screen-grid voltage.*

is secured provided the peak-to-peak amplitude of the f.m. signal at the grid is greater than the grid base. Solid-state equipment is arranged so that the limiter transistor is caused to "bottom" on normal signal amplitudes (see Chapter 5).

INTERFERENCE ALLEVIATION

Provided an efficient amplitude limiter is used in an f.m. receiver, the by-product f.m. caused by interfering pulses and signals is the only factor that will provoke a response at the loudspeaker and interference with the programme. Fortunately, the effect of interference on an f.m. receiver is less severe than on an a.m. one as the result of noise pulses or heterodynes of equal magnitudes.

To illustrate this point, consider the effect of interference on a.m. and f.m. systems both having equal transmission bandwidths—say 15 kHz (that is, an overall pass-band of 30 kHz). This means that maximum deviation of the f.m. system is limited to plus and minus 15 kHz. On the a.m. system the receiver

responds uniformly to interfering sidebands on the wanted signal over the whole of the audio spectrum (Fig. 2.8a). On the f.m. receiver, however, the noise output at 15 kHz is equal to the noise output of the a.m. receiver at the same frequency, but—and this is the important point—the noise response diminishes

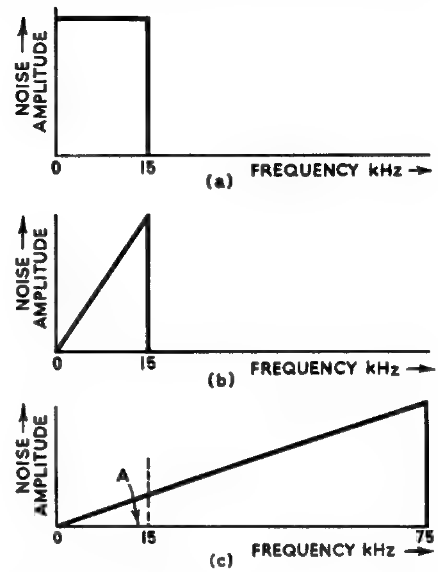


FIG. 2.8. *(a) How an a.m. receiver responds uniformly to interfering sidebands on the wanted signal over the whole of the audio spectrum. (b) The triangular noise characteristic of the f.m. system. (c) How the noise output is reduced by a further five times when the deviation is extended to plus and minus 75 kHz.*

proportionally to zero as the frequency of the noise components decreases from the maximum of 15 kHz to zero. In other words, the noise output increases in proportion to the frequency of the noise components or sidebands. This is brought about because the f.m. receiver responds in proportion to the frequency *difference* between the wanted carrier wave and the noise sidebands. The effect produces a triangular noise characteristic as shown in Fig. 2.8 (b).

When making a comparison of the noise output of the two systems in this way it is assumed that both sets have a linear a.f. response to the limit of the transmission bandwidth (15 kHz in the case under consideration) and that the a.m. system is modulated to a depth of 100 per cent. This latter stipulation must be made because with the f.m. system we are considering a *maximum* deviation of plus and minus 15 kHz—remembering that deviation is synonymous with depth of modulation.

The outcome of this is that the f.m. system gives a 2 to 1 better signal-to-noise performance when operating under similar conditions to an equivalent a.m. system. In regard to general circuit and valve or transistor noise the ratio is not quite as large and is reduced to something like 1·7 to 1 because of the random nature of the phase relationships between the carrier wave and the noise components at their various frequencies.

The maximum deviation of the current system of f.m. broadcasting is set at plus and minus 75 kHz; the effect that this has on the noise output of an f.m. receiver will now be considered. In order to cater for the extended deviation the bandwidth of the receiver has to be widened correspondingly, so a noise output versus frequency diagram can now be drawn as in Fig. 2.8 (c). There is a much

larger area of noise in this diagram than in the other two, but it will be realized that the noise components above about 15 kHz will produce heterodyne frequencies in the supersonic region—above normal hearing—and are therefore of little consequence. It can be seen that the audible interference is concentrated in the area A, and this is even smaller than the noise area of diagram (b).

What has happened is that the deviation has been increased five times, from 15 kHz to 75kHz, and this has increased the signal output five times; so by bringing the comparison back to the original basis, as from diagram (b) to diagram (c), the noise output is reduced in the same ratio. Thus, by increasing the deviation to plus and minus 75 kHz the signal-to-noise ratio is improved by a further five times, on the basis of an a.f. pass-band of 15 kHz.

PRE-EMPHASIS AND DE-EMPHASIS

With amplitude modulation the signal-to-noise ratio of the system cannot be improved by putting up the modulation percentage, which is really what has happened on the f.m. system by increasing the deviation, since we are limited to 100 per cent modulation and above this severe distortion results.

Because with f.m. most of the noise is introduced at the higher audio frequencies, it is feasible to enhance the signal-to-noise ratio still further by the use of pre-emphasis at the transmitter and corresponding de-emphasis at the receiver. At the transmitter the modulation is arranged to have a rising frequency characteristic giving greater amplification of the higher modulation frequencies than the amplification given to the lower ones. At the receiver the converse effect is achieved—the a.f. circuits are arranged to compensate for this rising-frequency characteristic by amplifying the lower modulation frequencies to a greater degree than the higher ones. In this way the noise which accompanies the higher modulation frequencies is correspondingly reduced without detracting from the quality of reproduction.

The audio frequencies above about 4,000 Hz are usually the harmonics which provide the true timbre to the instrument or voice. Since the amplitude of these harmonics is usually considerably below the amplitude of the fundamental frequencies, there is little fear of them over-modulating the transmitter, even when they are boosted in relation to the fundamental frequencies.

Frequency-modulation pre-emphasis is very much like that given to recordings. On disc recordings pre-emphasis is applied around 2 kHz; that is, the stylus velocity is increased with frequency above the figures mentioned as a means of increasing the signal-to-noise ratio on playback. Here also it will be appreciated that the lower energy content of the higher audio frequencies is often masked by the surface noise of the record. Thus, at the playback amplifier the extended top can be reduced to normal by a top-cutting circuit, which reduces the worst of the surface noise at the same time.

It may seem strange that the amount of de-emphasis and pre-emphasis is measured in time (microseconds). In the British f.m. system the pre-emphasis is 50 μs; it is 75 μs in America. This means that the time-constant of the de-emphasis circuit in the receiver should be 50 microseconds to restore the frequency characteristic of the modulation signal to the correct relative level.

The resistor-capacitor combination shown in Fig. 2.9 (a) represents a simple de-emphasis circuit which has a time-constant equal to C times R. The correct time-constant of 50 microseconds is thus secured by making R, say, 50,000 ohms and C 0·001 microfarad. Different values could be substituted provided their product equals 50. Some commercial receivers and tuner units, however,

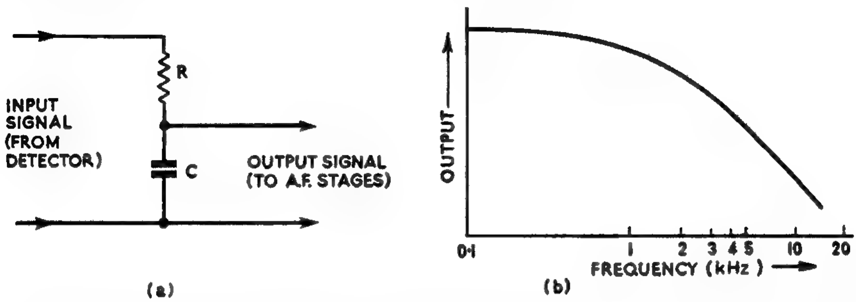


FIG. 2.9. *(a) Simple de-emphasis circuit. (b) The curve shows how the output voltage falls with frequency.*

may seem to have values in the de-emphasis circuit which do not work out to 50 microseconds. This is often because additional capacitive loading is contributed by other sections of the circuit, such as screened leads and coupling networks. For example, the signal-output lead of a number of f.m. tuner units is invariably screened to prevent the pick-up of hum, and since the capacitance of this screened lead is virtually in parallel with the de-emphasis capacitor, the value of the capacitor is reduced so that the total capacitance works out to that required to provide the correct degree of de-emphasis.

Some receiver designers purposely deviate from the required values as a means of boosting the bass or over-emphasizing the top to compensate for the non-linear frequency characteristics of the a.f. section of the receiver.

Returning to Fig. 2.9 (a), it will be seen that the input signal is applied across the resistor and capacitor in series and that the output signal is taken from across the capacitor. This network looks to the a.f. signal voltage as a potential-divider whose capacitive section C is in terms of capacitive reactance (X_c). Therefore, at very low audio frequencies the value of X_c (which, in ohms, is equal to $1/2\pi fC$, where f is the frequency of the input signal and C is the value of the capacitor in farads) is high in relation to the value of R, so nearly the full input voltage is passed to the a.f. stages of the receiver.

As the frequency of the input signal rises, however, the value of X_c decreases and the output voltage is progressively reduced. The output voltage falls with increase of frequency as shown by the curve in Fig. 2.9 (b). Usually, the loss of the network at various frequencies is expressed in decibels (dB). The turnover frequency occurs when X_c equals R.

At 15 kHz, which has been considered to be the limit of normal audibility, the further reduction of noise due to the processes of pre-emphasis and de-emphasis amounts to something like five times. It will now be realized that the cumulative effect of the noise-reducing artifices possible with the f.m. system is considerable when comparison is made with the a.m. system. Theoretically, the

noise output of a wide-band f.m. system with an audio pass-band of about 15 kHz is approximately only one twenty-eighth of the noise output of an a.m. system having the same audio pass-band.

Clearly, the main benefit of f.m. over a.m. is the vast improvement in the signal-to-noise ratio. This greater freedom from noise and interference generally is really an aid to a better quality of reproduction, even though the f.m. system as a whole possesses no inherent powers to secure it by other means. Freedom from noise and interference from other stations is possible without turning on the so-called tone control or brilliance control, so detracting from the quality of reproduction. The f.m. system also facilitates an extension of *dynamic range*, which is a hi-fi attribute.

CHAPTER 3

F.M. Detectors

IT has already been seen that in order to extract the audio-frequency components of the frequency-modulated signal for application to the receiver's a.f. and output stages a circuit is required the make-up of which is entirely different from that of the ordinary a.m. detector. What is required is a detector which responds to frequency variation, as distinct from the a.m. detector which responds to variation of amplitude of the carrier wave.

In Chapter 1 (Fig. 1.14) it was shown how the simple artifice of detuning can be used to give frequency to amplitude conversion. The a.m. output derived by this means will be handled by the receiver in the usual manner, and will produce an a.f. output from the loudspeaker in sympathy with the applied modulation.

Although this simple method outlines the basic function of an f.m. detector, its practical application is most inefficient and is definitely not a recommended way of detecting f.m. signals. Long ago, a similar arrangement was used in the American Fremodyne receiver, but it was soon outmoded in favour of the more efficient f.m. detector.

Severe distortion is possible owing to the operation of the scheme being focused on one side of the response curve of a tuned circuit which is liable to be far from linear. Moreover, the advantage of interference and noise suppression is not fully realized, and the resulting a.f. output signal is small compared with what it would be from an equivalent a.m. signal. The method can only be considered suitable for use in cheap receivers where low cost is considered to be of more importance than quality.

Nevertheless, consideration of the system reveals that the f.m. detector is called upon first to perform the function of converting the f.m. carrier to an equivalent a.m. one, and then the function of detecting this in the usual way.

FREQUENCY-AMPLITUDE CONVERTER

During the course of development of f.m. a large number of specialized detector circuits have been evolved which in turn have been susperseded progressively by arrangements of modern mode more suited to the demands of the time. A successful f.m. detector circuit evolved in the early 1920s, and known as the Travis frequency-amplitude converter, is shown in Fig. 3.1.

The circuit is very much like that of two ordinary a.m. diode-detector circuits arranged in 'push-pull'. In effect, this is really all the circuit is, but instead of the

two diode anode circuits being tuned to the same frequency, they are slightly off-set.

The primary winding L1 of the i.f. transformer, usually called the discriminator transformer, is in the anode circuit of the final i.f. amplifier valve which is

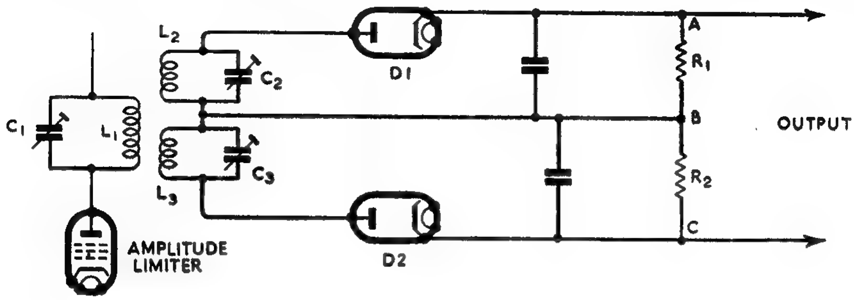


FIG. 3.1. *The Travis frequency-amplitude converter.*

also acting as an amplitude limiter. The primary is tuned by C1 to the unmodulated carrier frequency which, for the sake of discussion, let us suppose is 10 MHz. The two tuned circuits L2, C2 and L3, C3 form the secondary winding of the discriminator transformer. One is tuned above and the other an equal amount below the unmodulated carrier frequency. Usually, they are tuned to something like plus and minus 100 kHz from the carrier frequency so that they will respond properly over the full deviation of the f.m. signal.

Although the tuned frequencies of the two secondary circuits are displaced some 100 kHz above and below that of the unmodulated carrier frequency, the damping of the circuits is such that they respond equally to the unmodulated signal. In other words, the carrier frequency cuts the response curves of the two circuits at the same point, somewhere towards the bottom of the curves, 100 kHz off resonance.

This means that the application of an unmodulated carrier will cause equal currents to flow in the two diodes, D1 and D2. In addition, the voltages developed across the associated load resistors R1 and R2 will be equal as they are of equal value. The voltage developed at the cathode of D1 is positive with respect to point B, and the voltage developed at the cathode of D2 is positive with respect to point B. The voltages are thus in opposition and, because they are equal, give zero voltage between points A and C. This is the condition of balance when the signal is unmodulated.

Let it be supposed that in the course of modulation the instantaneous frequency of the carrier swings up the side of the response curve of L2, C2 towards its resonant frequency. The current in D1 rises and a rising voltage occurs across the load resistor R1. It follows that since the instantaneous frequency of the carrier has shifted towards the resonant frequency of L2, C2 it must have moved down the side of the response curve of L3, C3 away from its resonant frequency, thereby causing a reduction of current in D2 and a reduced voltage across the load resistor R2.

As an illustration, it is feasible that the instantaneous swing of the carrier

frequency caused a one-volt increase across R1 and a one-volt decrease across R2, thereby giving a net rise of two volts from zero (from the unmodulated condition) between the output points A and C. Clearly, the variations of voltage across the two resistors occur in sympathy with the frequency variations of the carrier wave; they thus represent the modulation frequency, which can be conveyed to the a.f. stages for amplification and ultimate reproduction.

It can be seen from Fig. 3.2 that the action of the system is similar to that of the detuning method. As opposed to the simple detuning method, however,

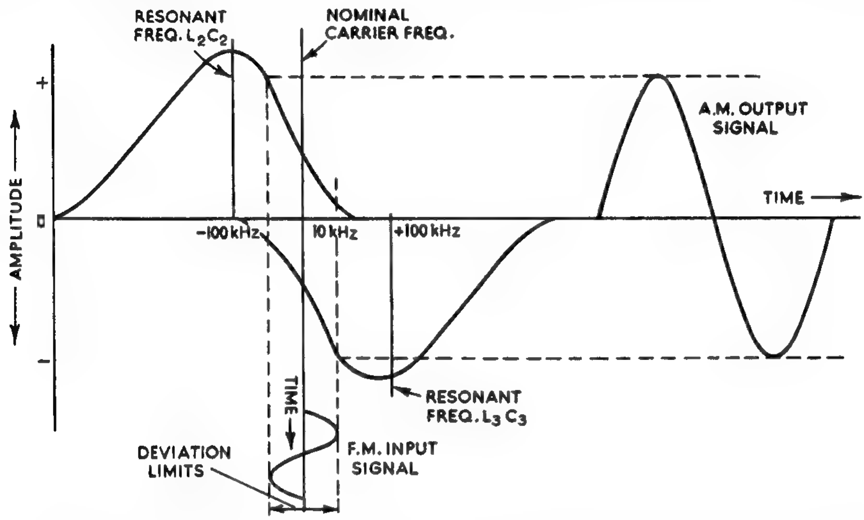


FIG. 3.2. *Curves illustrating the function of the Travis frequency-amplitude converter. This is something like the detuning arrangement in Fig. 1.14.*

the two-diode circuit is considerably more symmetrical and, provided the responses of the two secondary tuned circuits are perfectly matched and equally off-tune in relation to the carrier frequency, very little non-linear distortion results.

Of academic interest is the frequency-amplitude converter which features triode valves instead of diodes. From the circuit in Fig. 3.3 it will be observed that it has much in common with the Travis arrangement.

The two triodes are connected as super-regenerative detectors. This means that the valves are connected in relation to the tuned circuits L2, C2 and L3, C3 so that they tend to go into heavy oscillation. Instead of the oscillation being continuous, however, it is quenched at regular intervals by the potential of the grid capacitors C4 and C5 charging up and pushing the valves into cut-off. The time constants of C4, R1 and C5, R2 are such that the quenching frequency is held constant beyond the audio range and is not heard from the loudspeaker.

A remarkable increase in sensitivity over the diode method is possible with the super-regenerative system, much in comparison with the ordinary diode or crystal a.m. detector and the leaky-grid detector with reaction. Unfortunately, the circuit is not easy to adjust properly; it tends to be considerably unstable and it introduces a.f. distortion.

From the f.m. aspect, the system functions after the style of the Travis arrangement; the primary of the transformer L1 is tuned by C1 to the unmodulated frequency of the signal, and the tuned circuits L2, C2 and L3, C3 are tuned

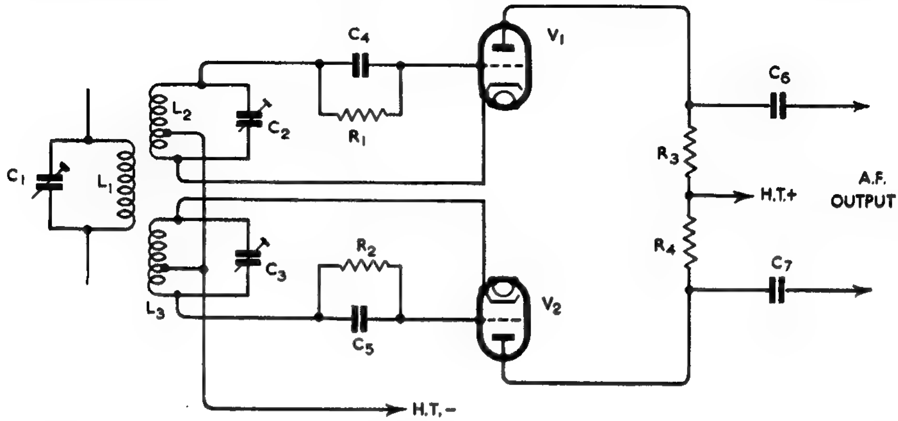


FIG. 3.3. *A super-regenerative frequency-amplitude converter. This circuit provides a remarkable increase in sensitivity over the two-diode circuit.*

above and below that of the carrier frequency in the manner already described. Because the valves are operating from between cut-off and saturation, due to the oscillatory and quenching actions, they provide automatic limiting. Clearly, amplitude fluctuations of the applied signal will push the valves further into cut-off or else further into saturation and will, therefore, have little or no effect on the output signal. The a.f. signal is developed across the anode resistors R3 and R4 and conveyed to the a.f. stages through capacitors C6 and C7.

THE PHASE DISCRIMINATOR

With the Foster-Seeley circuit the resulting phase displacement of two signal voltages—produced by the f.m. signal deviating above and below that of the resonant frequency of a tuned circuit—causes the production of voltage variations in sympathy with the modulation signal. This statement is not so complex as it sounds; it simply means that, instead of the a.f. being derived from changes in voltage as the applied f.m. signal swings up and down the sides of the response curves of two tuned circuits (as in the previously considered examples), the output voltage is obtained by reason of phase changes as the f.m. signal swings between its deviation limits. For this reason the Foster-Seeley circuit is often referred to as a *phase discriminator*.

From the basic circuit in Fig. 3.4 it will be seen to have much in common with the two-diode circuit. The main differences are: (1) the inclusion of the coupling capacitor between the primary and secondary circuits of the discriminator transformer T1, and (2) the secondary is formed of one tuned circuit instead of two separately tuned circuits. As is common with most f.m. detectors, a connexion is made at the centre of the secondary winding.

Although the circuit looks simple enough, its exhaustive analysis is quite involved and demands the abundant use of mathematics such as are beyond the

scope of a book of this nature. Nevertheless, since the present purpose is not wholly concerned with the design of circuits, sufficient understanding can be secured by viewing the arrangement from a more practical aspect.

Reverting to the circuit in Fig. 3.4; both the primary and secondary of the discriminator transformer T1 are tuned to the intermediate frequency corresponding to the unmodulated signal, say, 10 MHz. A signal of this frequency is, therefore, produced by the i.f. amplifier valve V1 and developed across the primary and a voltage is induced, by transformer action, across the secondary. Incidentally, V1 is also probably arranged as an amplitude limiter.

Now, because the secondary has a centre-tap which is connected back to a common point in the cathode circuit of the diodes D1 and D2, each diode will

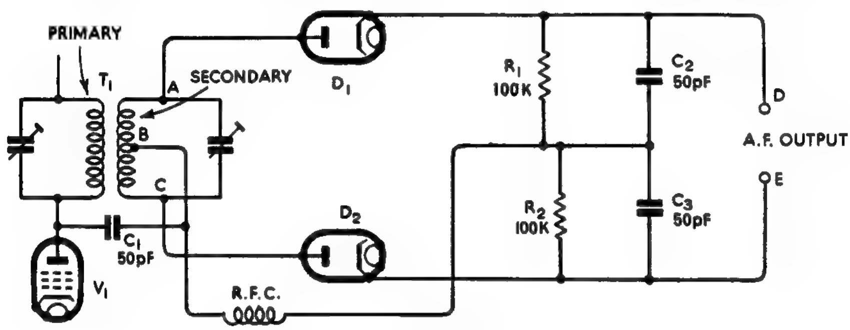


FIG. 3.4. *The Foster-Seeley phase discriminator.*

receive a signal of equal voltage, but a phase displacement of half a cycle (180 deg.) will exist between the two anode signals. It will be possible to appreciate this condition from the discussion on phase relationships in Chapter 2. The condition is equivalent to that at the anode side of a full-wave h.t. rectifier circuit. As one anode swings positive so the other swings negative by the same amount.

The diode load resistors R1 and R2 are also of equal value. The diodes therefore conduct equally and alternately, and since the resulting currents flowing in the two load resistors are equal, the voltages developed across them are equal but of opposite sign. This part of the operation is the same as with the Travis frequency-amplitude converter—both cathodes go positive relative to the junction of R1 and R2, which is negative to both cathodes. The voltages across the resistors are added in opposition across the output terminals D and E and resolve so as to provide zero output.

So far all is well, but from the present consideration the circuit is pointless since, even if the signal deviates from its nominal unmodulated frequency, the diodes would still conduct equally and there would still be zero output across points D and E.

There are two other factors to be considered. One is the coupling capacitor C1; this is connected from the anode V1 to the centre-tap on the secondary. The signal across the primary is conveyed by this capacitor to the centre-tap so that it is applied, together with the induced signal voltage in the secondary, to the two diodes.

The conditions now are that diode D1 receives the signal voltage across the points AB of the secondary plus the signal voltage contributed by way of C1, and that diode D2 receives the voltage across BC plus the signal voltage contributed by C1.

THE PHASE ASPECT

At this stage it is necessary to have some knowledge of an important property of tuned and coupled circuits. It is that in an r.f. transformer having both primary and secondary windings tuned to a particular frequency, there is a phase displacement of 90 deg. between the voltages across the two coils at the frequency to which they are tuned; and, more important—if the frequency differs from that to which the windings are tuned, this 90-deg. phase displacement no longer holds good, it either increases or diminishes depending on which way the coils are off tune in relation to the signal frequency. It is this property which makes the phase discriminator frequency-sensitive, and is the second factor to be considered.

When the primary and secondary of T1 (Fig. 3.4) are tuned exactly to the frequency of the signal, the signal contributed by way of C1 is 90 deg. out of phase with the signal induced across the secondary. This means that the anode of D1 receives a voltage from section AB of the secondary plus a voltage—differing in phase by 90 deg. from that of the primary—through C1. Similarly, the anode of D2 receives a voltage from section BC plus a voltage—also differing in phase by 90 deg. from that of the primary—through C1.

The conditions now, with respect to the centre-tap, are that the voltage across the primary, and the voltage across BC is 90 deg. either behind or in front of the voltage across the primary. Remember, there is a phase difference of 180 deg. between the two diode anodes, or points A and C.

Since the phase difference between the two voltages at the two diodes is 90 deg., the voltages at the diode anodes will be equal, no matter that the 90-deg. displacement is in the opposite sense at each anode.

In Fig. 3.5 are shown vectors representing the voltages across the various points in Fig. 3.4 when the signal frequency equals the tuned frequency of the

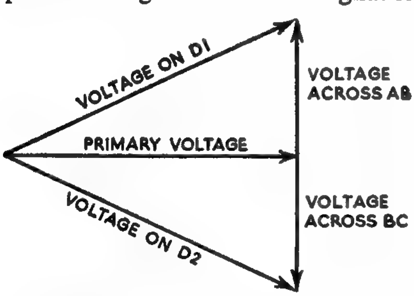


FIG. 3.5. *Vector diagram showing the condition of balance in a discriminator.*

transformer. It can be seen that the voltage across AB is 180 deg. out of phase with the voltage across BC, and that the voltage across the primary is 90 deg. out of phase with the voltage across AB and 90 deg. out of phase, in the opposite direction, with the voltage across BC.

By adding vectorially the voltage across AB and the primary voltage, and the voltage across BC and the primary voltage, it will be seen that the voltage on

D1 is equal to the voltage on D2. This is the condition of balance when the signal frequency equals the tuned frequency of the transformer; this causes the two diodes to conduct equally and thus give zero output across the two series-connected load resistors.

Consider what happens when the signal changes in frequency when it is frequency-modulated. Let it be supposed that in the course of modulation the instantaneous frequency of the carrier swings, say, 50 kHz away from the tuned frequency of the transformer. The 90-deg. phase displacement no longer holds. The voltage across one half of the secondary swings towards the phase of the voltage across the primary, and the voltage across the other half swings away from the phase of the voltage across the primary.

This condition is shown by the vector diagram in Fig. 3.6, which also reveals that the balance of the voltages at the diodes is considerably disturbed. From

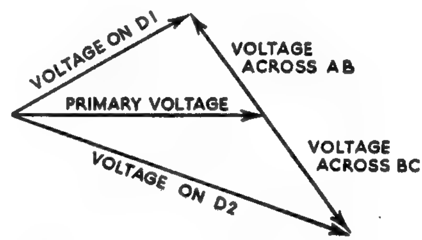


FIG. 3.6. *Vector diagram showing the condition of unbalance in a discriminator.*

the illustration, the voltage on D1 has decreased and the voltage on D2 has increased.

The vectors show that, although the magnitude of the voltages concerned remains the same, the output due to two voltages differing in phase changes as the phase angle changes. When two in-phase voltages are added together the output is equal to the sum of the two voltages, and when the voltages are 180 deg. anti-phase they are added vectorially; that is, plus or minus, the lesser being subtracted from the greater. This means that if the voltages are equal and anti-phase they will cancel out and the output will be zero. Between the in-phase and anti-phase conditions, therefore, the output will vary between the sum and difference of the two voltages.

A condition between in-phase and anti-phase is shown in Fig. 3.6. It is interesting to note that the angle of tilt away from the 90-deg. state is proportional to the depth of modulation (deviation), while the rate at which the angle changes is representative of the modulation frequency—it being remembered to keep these factors in their proper places.

It will now be possible to realize that as the frequency of the carrier rises and falls, the balance of the circuit is disturbed in rhythm with the modulation, and the a.f. signal is developed across the series-connected load resistors R1, R2 of Fig. 3.4.

Capacitors C2 and C3 serve to rid the a.f. section of any residual r.f. voltage, while the radio-frequency choke (r.f.c., Fig. 3.4) is included in the centre-tap circuit to prevent the primary of T1 from being damped by the cathode circuit of the diodes, and also to aid in blocking the signal at the primary, via C1, from the a.f. section.

Provided the characteristics of the discriminator transformer are such that at peak deviation the phase displacement between the primary and secondary does not exceed about 45 deg. the circuit provides a remarkably distortion-free output but, in common with all other discriminator circuits, optimum results

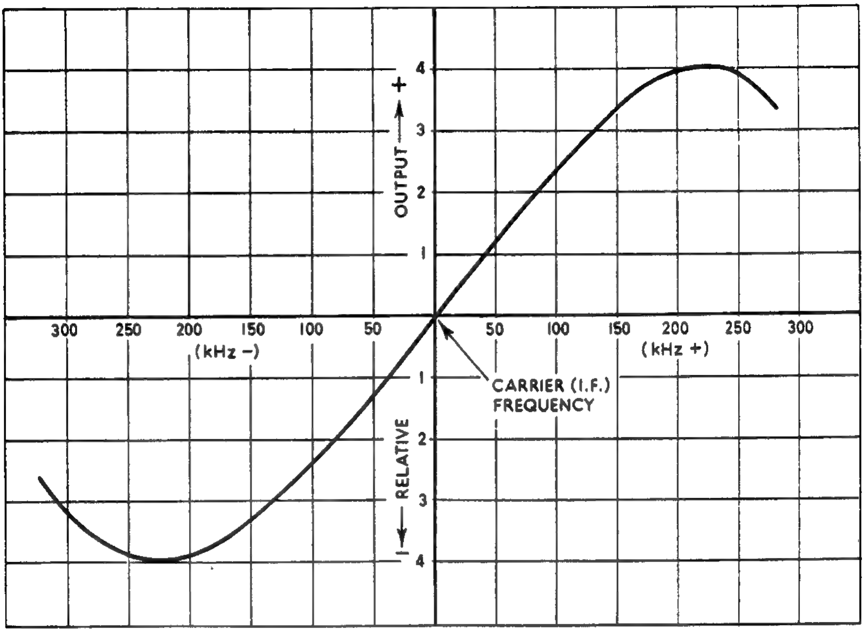


FIG. 3.7. *A typical phase-discriminator characteristic.*

are possible only when the diode circuits are perfectly balanced in relation to the centre-tap on the secondary and when the transformer is properly aligned in relation to the i.f. stages.

A typical discriminator characteristic is shown in Fig. 3.7. It will be seen that only the most linear part of the curve is used for a deviation of plus and minus 75 kHz.

THE RATIO DETECTOR

It would seem from modern circuit design that the ratio detector is the most favoured of all f.m. discriminators. This is most certainly due partly to the fact that the ratio detector is endowed with a limiting feature which tends to preclude the passage of a.m. disturbances of the signal. It will be seen later how this is achieved. For the present it is necessary to have some idea of the operation of the circuit.

In Fig. 3.8 is shown a typical ratio-detector circuit. This differs from the previously described discriminator in that the diodes V2 and V3 are connected in series—instead of back-to-back—and a small tertiary winding L3 is used in place of a capacitor to transmit a phase difference from the primary to the secondary of the discriminator transformer.

It is important to note that the voltage across the tertiary winding is always in phase with the voltage across L1. This is because the tertiary winding is

virtually untuned and possesses a resonant frequency far removed from the tuned frequency of the primary and secondary of the discriminator transformer L1 and L2.

As with the Foster-Seeley circuit, when the frequency of the applied signal is equal to the tuned frequency of the transformer (this represents the condition of balance) each diode will receive an equal but opposite voltage. In this circuit also it will be seen that L2 is centre-tapped and that this circuit is returned to a point

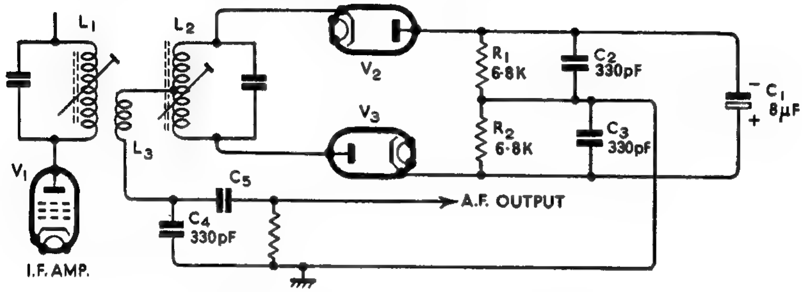


FIG. 3.8. *Circuit of a typical ratio detector.*

of balance in the diode circuit. Under the conditions of balance, therefore, the currents in the diode circuits will be equal, and since the diodes are connected in series a steady direct voltage will develop across R1, R2, and the capacitor C1 will acquire a charge.

When the signal is frequency-modulated the voltage across L3 adds to the voltage across one side of L2 and subtracts from the voltage across the other half, in the same way as with the Foster-Seeley circuit. The diode receiving the larger voltage will conduct more heavily than the other, but the diode receiving the smaller voltage will not be able to pass this extra current because the two diodes are series-connected; this extra current flows out of the series diode circuit via L3, C4 and then back at the junction of R1, R2.

This may be regarded in another way. The diodes V2 and V3 can be considered as variable resistors whose values are governed by the voltages applied to them—when the voltages are equal their values are equal. Since they are connected in a balanced circuit brought about by the centre-tap on L2 and the balanced load resistors R1 and R2, it follows that the voltage between the tap on L3 and the junction of R1, R2 will be zero, provided the representative resistance values of V2 and V3 are equal. This can be considered as the condition of balance.

Now, supposing V2 to be reduced in value (as the result of an increase in voltage) and V3 to be increased in value (as the result of a decrease in voltage), then the balance of the circuit would be disturbed and current would flow out of the circuit to give rise to a voltage across the two points mentioned—much after the style of a bridge circuit.

In an earlier paragraph reference was made to the current flowing out of the circuit by way of L3 and C4. It may be wondered how current can flow through a capacitor, but it will be remembered that the out-of-balance current is occurring at audio frequency in sympathy with the audio modulation of the signal.

The out-of-balance current is brought about by unbalance of the diodes due to phase changes between the voltages across the primary and secondary of the discriminator transformer as the applied signal is frequency-modulated. Thus, the out-of-balance current in C4 produces an a.f. voltage across it, and this voltage is fed to the a.f. stages of the receiver through C5.

AMPLITUDE LIMITING

The voltage developed across R1 and R2 in series, due to the series diode current, charges capacitor C1 to a value equal to the voltage across R1 plus the voltage across R2. A continuous rectifying action of the signal thus takes place due to the series-connected diodes, and the magnitude of charge of C1 is proportional to the amplitude of the signal at the input of the detector.

As the amplitude of the signal rises so the voltage across C1 increases and, conversely, as the amplitude of the signal falls so the voltage across C1 decreases. The sum of the resistances of R1 and R2 and the value of C1 form a time-constant (CR). The value of this time-constant, usually in the region of 0·2 second, is arranged so that any rapid variation of the amplitude of the input signal, such as that due to impulsive interference and audio frequency, finishes before it can reveal its presence by modifying the average charge on C1.

Moreover, when the detector receives a signal of constant amplitude the voltage across C1 settles down to equal that across R1 and R2 in series. When the amplitude of the input signal increases, the voltage across R1 and R2 also increases and current flows into C1 to increase its charge. This flow of current looks to the diode circuit as a decrease in the value of the load resistance (R1 plus R2) and reflects greater damping on the transformer circuits. This tends to offset the increase of amplitude of the input signal and consequently has a stabilizing effect on the output voltage.

When the amplitude of the input signal decreases, the voltage across R1 and R2 also decreases and current flows out of C1 via the load resistors. The flow of current in this direction looks to the diode circuit as an increase in the value of the load resistance and reflects reduced damping across the transformer circuits, which again has a stabilizing effect on the output voltage.

It should be remembered that if an amplitude increase occurs then a larger voltage is applied to both diodes and the increase to each is equal. Their conduction currents are increased equally, but the out-of-balance condition is not altered and the a.f. output signal does not increase since it is not directly related to the amplitude of the signal above the minimum necessary to work the circuit properly.

THE UNBALANCED CIRCUIT

There are two types of ratio-detector circuit in common use. One of these is the balanced circuit which has already been considered. The other one is the unbalanced circuit shown in Fig. 3.9. Here, only one resistor R1 occupies the position of diode load and one side of this is taken to the chassis connexion instead of to the junction of two equal value resistors as in Fig. 3.8.

The unbalancing of the circuit in this way would appear to detract from the

desirable feature of a.m. rejection; this is true, but not to such a large degree as would be expected. The unbalance effect in this connexion can be considerably offset by increasing the value of the stabilizing capacitor C1. In practice this

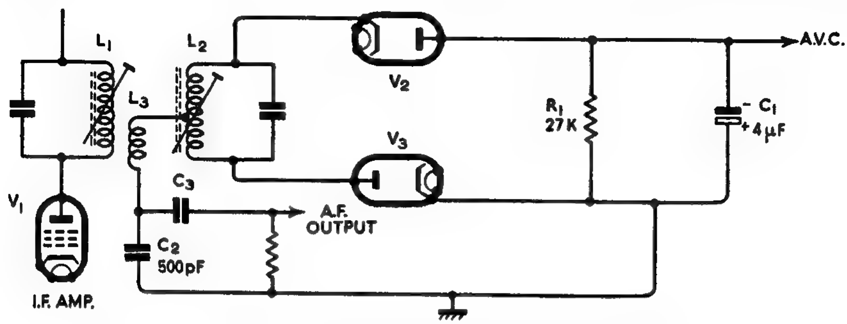


FIG. 3.9. *Circuit of unbalanced ratio detector. Compare this with Fig. 3.8.*

value is usually in the region of 4 to 8 microfarads, and the capacitor is of electrolytic type. A resistor of about 50,000 to 100,000 ohms is usually employed to provide the time-constant.

Increasing the time-constant by increasing the value of the stabilizing capacitor considerably aids in providing enhanced amplitude rejection when the frequency of the disturbance is relatively low. It also has the effect of making quick and accurate tuning of the receiver rather difficult. It is found necessary first to adjust the tuning of the receiver approximately so as charge the stabilizing capacitor which, if of a relatively large value, may take two seconds or so. After the capacitor has been initially charged by the signal the receiver can be tuned more accurately.

ADVANTAGE OF THE UNBALANCED CIRCUIT

For the sake of saving two capacitors (C2 and C3 in Fig. 3.8) and a resistor it may be considered an unwise and false economy not to take advantage of the balanced arrangement. However, the unbalanced arrangement makes it easier to obtain a source of voltage of suitable magnitude and polarity for controlling the gain of the r.f. and i.f. valves so that the receiver can possess the feature of automatic gain control (a.g.c.).

The voltage across the load resistor, as already discovered, is proportional to the amplitude of the input signal. With the balanced circuit only half the total voltage across the two load resistors, relative to chassis, is available for this purpose. Usually, this is too small and more voltage is required to work the a.g.c. circuit properly. With the unbalanced arrangement the full voltage is available across R1 with respect to chassis, and if the voltage is taken from the anode of one of the diodes (Fig. 3.9) it is negative and of a suitable level fully to control the first-stage valves. There are variations of the circuits.

METHODS OF ACHIEVING A BETTER BALANCE

As with most discriminator circuits, the better the balance of the circuit the better will be the rejection of spurious a.m. disturbances. Several artifices are

adopted for reducing the unbalanced a.m. component. One is to modify the tap on the secondary winding of the discriminator transformer so that in relation to the other circuit constants an almost perfect balance is achieved.

Another method is shown in Fig. 3.10 with reference to the balanced type of circuit; here, three additional resistors have been incorporated in the circuit.

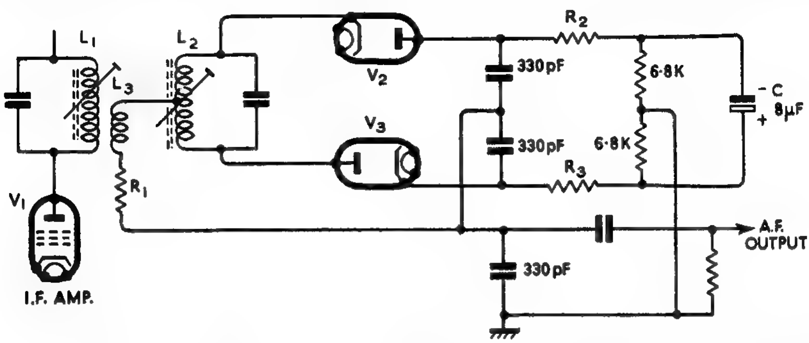


FIG. 3.10. *Enhanced rejection of a.m. disturbances can often be secured by introducing resistors R1, R2 and R3 as explained in the text.*

The resistor R1 limits the peak diode currents and aids in maintaining balance when the input signal is large. Resistors R2 and R3 introduce a deliberate unbalance as a means of countering unbalance in the circuit proper. The value of R1 is usually very low, in the region of 47 ohms, while R2 and R3 are unequal; one might be, say, 1,000 ohms and the other 1,500 ohms depending on how the circuit requires correcting; however, the introduction of extra resistors reduces the overall sensitivity of the stage.

USE OF SEMICONDUCTOR DIODES

It is not necessary to use valve diodes in discriminator circuits. Most f.m. equipment now makes use of semiconductor diodes, germanium or silicon, instead of thermionic valve diodes. These have the advantages that they are small and do not require heater current. They are often tucked away within the screening can of the discriminator transformer together with other associated components.

When employing semiconductor diodes, though, it is as well to ensure that they are matched before wiring them into the circuit and fitting them in the can of the discriminator transformer. A pair having equal forward resistances, as measured on the "ohms" range of a multi-range test meter, should be selected. Although this is only a rough check for balance it is better than making no test at all.

COUNTER-DIODE FREQUENCY DETECTOR

The counter-diode frequency detector, or the pulse-counting discriminator as it is sometimes called, uses a principle entirely different from the circuits so far considered. The arrangement has not been exploited to such a large extent commercially as the Foster-Seeley discriminator and the ratio detector, though it is understood to be used by home constructors in view of its simplicity of adjustment and extremely low distortion characteristic.

Principle of Operation. The number of cycles of an f.m. signal applied to the detector is proportional to the modulation frequency. If the signal is simply rectified, the mean value of the rectified signal will remain constant if the amplitude of the signal is constant, even though the half-wave pulses due to the rectifying action change in frequency with the modulation. This is because each

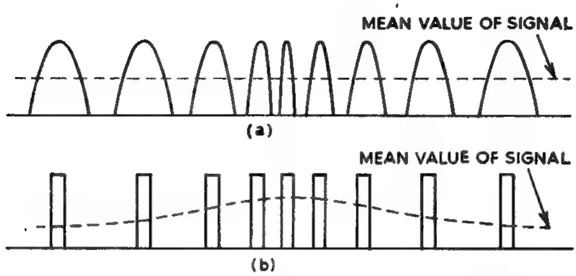


FIG. 3.11. *Showing at (a) that the mean value of the rectified signal will remain constant if the amplitude of the signal is constant, even though the pulses change in frequency with modulation. If the signal pulses are turned into square waves of equal size, as at (b), the mean value of the pulse chain will vary in accordance with its frequency variation.*

increase or decrease in the number of cycles is exactly balanced by their decreased or increased duration (Fig. 3.11a).

With the pulse-counting discriminator the half-wave pulses of the rectified f.m. signal are made all the same size, regardless of the frequency and amplitude of the signal. Actually, they are turned into somewhat distorted square waves which close up and open out with the modulation signal.

This effect is depicted in Fig. 3.11 (b) for the same signal as at (a). Now, when this is done, and the pulses are all the same size, the mean value of the pulse

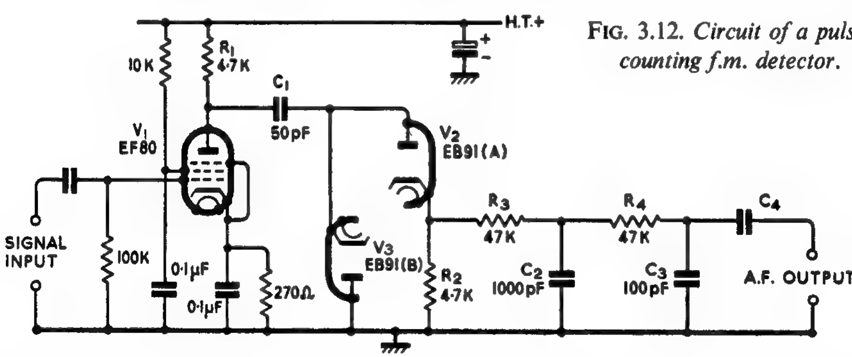


FIG. 3.12. *Circuit of a pulse-counting f.m. detector.*

chain will vary in accordance with its frequency variation, and if the pulse chain is derived from an f.m. signal the mean value will vary in accordance with the applied modulation—this, of course, being the fundamental requirement of an f.m. detector.

Fig. 3.12 shows a circuit of this kind of f.m. detector. The pentode valve V1 serves to produce square wave pulses from the f.m. signal applied to the grid.

The circuit constants are chosen so that the input signal causes the valve to operate from between saturation and cut-off.

The action of the circuit can be understood by imagining, for the time being, diode V3 to be disconnected from the circuit and diode V2 short-circuited. When the valve is driven into saturation the current in the anode load R1 is maximum and the voltage at the anode is minimum. The capacitor C1 thus develops a charge via R1 and R2. The voltage across C1 equals that at the anode, and when the charge is completed there is no current in R2 and so no voltage across it.

When the signal cycle changes and the valve cuts off, the current in R1 falls to zero and the voltage at the anode rises to that of h.t. positive. Since C1 cannot follow this voltage rise instantaneously, though, as the result of the time-constant (R1 + R2)C1, the increase in voltage is reflected across R1 and R2 and is distributed in proportion to their values. This results in the creation of a sharp rise of voltage across R2, which decreases exponentially as C1 gradually charges to the value of the anode voltage. A positive-going pulse is thus produced on that particular signal cycle.

If the action of the circuit is still considered without the function of the respective diodes, a negative-going pulse of similar form would be produced across R2 when the signal cycle changes again and the valve is driven into saturation. Since we are interested only in positive-going pulses, however, diode V2 serves to cut off R2 from negative voltages, and diode V3 takes over to discharge C1.

CONSTANT PULSE SIZE

Provided the circuit time-constant and the voltage of the pulses are held constant, equal-sized pulses will be developed across R2, irrespective of frequency variations. The circuit time-constant obviously remains stable, and since V1 is alternating between saturation and cut-off its function is after the style of an amplitude limiter, so the pulses are also held at a constant voltage in spite of variations in signal amplitude.

Thus, the rectified voltage appearing across R2 follows the modulation of the f.m. signal. Components R3, C2 and R4, C3 form a two-stage filter of twofold function. It provides the required degree of de-emphasis and removes the r.f. component of the f.m. signal. The output is fed through C4 to the receiver a.f. stages.

LOW-VALUE INTERMEDIATE FREQUENCY

The time-constant of the circuit in relation to the frequency of the input signal is extremely critical. The time-constant must be arranged so that the pulse is almost completed before the valve changes its mode of operation. Disregard of this stipulation would result in distortion since the pulses would not all be exactly the same size and the mean value of the pulse chain would not be exactly proportional to frequency.

Moreover, for correct limiting action of the circuit, coupled with providing a reasonable output voltage, the values of R1, R2 and C1 must be computed

within certain limitations. The values so resolved govern the range of frequency of the input signal during the process of frequency-modulation. In general, it ranges between a maximum of 250 kHz and a minimum of 100 kHz for a deviation of plus and minus 75 kHz, thereby giving a nominal i.f. of 175 kHz.

The incoming f.m. signal can be reduced to this very low i.f. in one step by operating the local oscillator of the frequency changer accordingly, or else two frequency changers can be used, one to reduce the f.m. signal to, say, 10 MHz, at which frequency it can be amplified in the conventional manner, and another frequency changer to reduce the 10 MHz i.f. to 175 kHz for ultimate application to the pulse-counting circuit. Both methods have successfully been used in practice though for high sensitivity the double-frequency-changing method is usually demanded.

The advantages of the circuit are its low distortion output and the feature of needing no alignment. If the f.m. signal is reduced to the low i.f. in one step the "no alignment" feature can be extended to the i.f. amplifier stages, for these can be resistance-coupled and thus operate without tuned circuits. One disadvantage of this method is that the low i.f. will cause the second-channel response to fall very close to the main response, producing two tuning points of the same station separated by a little over 300 kHz.

Pre-set switch positions can resolve this problem, and complete accuracy in tuning can be ensured by crystal control of the local oscillator or by automatic frequency control (a.f.c.). Spurious signals falling in the i.f. pass-band do not usually cause interference owing to the capture effect.

THE NONODE F.M. DETECTOR

Another interesting form of f.m. detector which also features automatic amplitude limiting is shown in Fig. 3.13. The circuit is evolved round the multi-

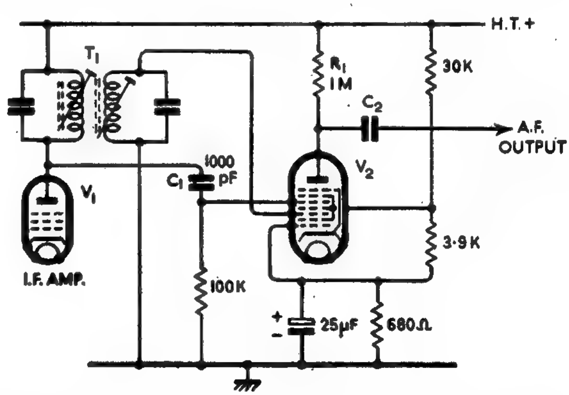


FIG. 3.13. *Circuit diagram of the nonode f.m. detector in which the anode current of V2 is arranged to vary in sympathy with the modulation pattern of the f.m. signal.*

grid valve V2, known as a nonode; V1 is the normal i.f. amplifier valve which is coupled to the nonode via the i.f. transformer T1.

Two voltages having a phase displacement between them of 90 deg. are applied to the third and fifth grids of V2. This phase displacement exists only as long as the frequency of the input signal equals the tuned frequency of the i.f. transformer, as with the phase discriminator and ratio detector. The signal at

the third grid is taken from the secondary of the i.f. transformer and the signal at the fifth grid is taken from the primary via C1.

The voltages at the first and second grids are held constant by the circuit and, as with a pentode valve, the electron current is likewise held constant as a result. Anode current flows only when the third and fifth grids swing positive. Since the magnitude of the voltages at these grids depends on the phase difference, which varies in accordance with the modulation of the f.m. signal, the anode current must vary in sympathy with the modulation. The a.f. signal is thus developed across the anode load resistor R1 and fed to the receiver a.f. stages by way of C2.

Because the electron current is held at a constant value the valve will not give an output as the result of amplitude variations of the input signal, though it is unfortunate that a relatively large f.m. signal is required to secure a proper amplitude-limiting action. A similar arrangement is used in some 625-line TV receivers.

CHOICE OF DETECTOR

Commercially, the ratio detector and the phase discriminator are the most favoured circuits. With combined a.m./f.m. receivers the ratio detector is invariably adopted mainly because of the ease in which it can be incorporated with the normal a.m. diode detector by the use of a common triple-diode-triode valve (such circuit arrangements are discussed later) and because it provides automatic amplitude-modulation rejection, thus rendering a separate limiter stage unnecessary. Where commercial combined receivers are concerned, economy of circuit structure is a big factor in favour of the ratio detector.

In combined receivers the ratio detector is nearly always of the unbalanced mode, thereby easily permitting an a.g.c. potential to be taken from across the stabilizing capacitor. This potential is also often used to work a magic-eye tuning indicator.

In f.m. tuner units which are designed for use with a high-fidelity a.f. amplifier or for connexion to the pick-up sockets of a radio receiver, the phase discriminator is sometimes featured. The same applies to the few receivers designed for f.m. use only.

It has been said that the Foster-Seeley circuit introduces slightly less distortion than the ratio detector at peak deviation, but in practice it is difficult to detect differences in distortion between the two systems even under favourable conditions when very high-quality a.f. equipment is employed. It must be remembered that the transmission is rarely at full deviation, so the distortion is correspondingly less. In addition, there are all the other distortion-producing agents in the complete f.m. transmitting-receiving chain to be taken into consideration when one becomes too critical from the technical aspect.

It is most important to remember that high distortion figures are possible unless the discriminator, or ratio detector, is properly designed and adjusted. The Foster-Seeley circuit is more critical in this respect than is the ratio detector and requires careful adjustment of the coupling between the primary and secondary windings of the discriminator transformer in order to keep the distortion content to a low figure.

As the home constructor is unlikely to possess equipment for measuring and

adjusting i.f. transformer coupling factors, it is desirable that he keeps strictly to the specifications given in the instructions for building the i.f. transformers, particularly the discriminator transformer. Where possible it is best to use the balanced mode when the ratio detector is adopted.

If very low distortion figures are required, coupled with comparative ease of design and adjustment, the pulse-counting circuit should be given due consideration. It is considerably less sensitive than are resonant discriminators, and if the f.m. signal is converted to the low i.f. in one stage, the circuit would hardly be of use in areas beyond the 1 mV per metre contour. There exists the possibility of using, say, three resistance-coupled i.f. stages and a high-gain a.f. stage following the pulse-producing circuit as a means of compensating for the inherently low sensitivity. It is unsuitable for stereo, however.

The nonode circuit also lacks sensitivity and for efficient limiting requires over 11 volts peak signal at the input grids. For those who wish to experiment with this circuit the nonode valve comes under the designation EQ80.

INTEGRATED CIRCUIT F.M. DETECTORS

Linear operational integrated circuit (i.c.) amplifiers designed for f.m. applications often incorporate diodes which can be used for the f.m. detector in addition to the bipolar transistor elements and resistors of the i.f. channel amplifier. These diodes can be arranged to form a fairly conventional ratio detector; but since it is not possible to build high-capacitance elements into ordinary i.c.s, the detector circuit is sometimes arranged to work without the usual stabilizing electrolytic across the load. Instead the capacitance of reverse-biased diodes within the i.c. is used in a manner to produce so-called "average detection".

Sometimes, though, the i.c.s are arranged essentially for i.f. signal amplification, separate external semiconductor diodes then being used in the detector circuit in the normal way, with the electrolytic across the d.c. load.

At the time of writing (1969) a number of i.c.s have been developed for f.m. i.f./detector applications (see, for example, the reference to the RCA i.c.s in Chapter 5), and one rather interesting scheme, which is likely to be adopted more extensively over the years, uses an i.c. with two pairs of transistor elements acting as one amplifier channel and two more transistor elements in a second channel.

The i.f. signal is amplified and limited in one channel to yield a rectangular output waveform (due to the limiting) and then split into the second channel which contains a tuned circuit (externally connected to the i.c.), the action of which restores the 10·7 MHz rectangular signal to normal sine-wave form at the output of that channel.

Both the rectangular waveform and the sine wave are fed to a "coincidence" type of detector, and since the modulation varies the instantaneous frequency of both signals, and since the sine wave in subjected to a phase displacement due to the action of the tuned circuit, the coincident detector produces an output consisting of a series of pulses of mean value proportional to the modulation frequency. The action of this detector is similar to the pulse-counting arrangement illustrated in Fig. 3.11.

There is, in fact, another i.c. (Intermetall Type TAA710) which incorporates a complete f.m. system, including the oscillator, mixer, i.f. amplifier and mixer, requiring a few external components, such as resistors and capacitors, to operate and to yield an a.f. output up to 1V maximum.

CHAPTER 4

R.F. and Frequency-Changer Stages

For the sake of continuity of presentation the discussion of f.m. detectors is followed by a consideration of the radio-frequency and frequency-changer stages, though this sequence is contrary to that of the signal itself.

Frequency-modulated sets differ essentially from their a.m. counterparts from the aspect of the detector and the v.h.f. signal-frequency stages. This applies equally to all modes of f.m. receiving equipment, embracing receivers designed solely for the reception of frequency-modulated signals, f.m. adaptors, combined a.m./f.m. receivers and combined television/f.m. receivers.

With combined receivers the process of conveying either the frequency-converted a.m. or f.m. signal to the appropriate detector stage and thence to the a.f. section is, of necessity, complicated (in comparison with f.m.-only receivers and f.m. adaptors) by switched circuits and additional circuit networks, but combined receivers will be considered in some detail in a later chapter.

In the very-high-frequency spectrum, the amplifying and frequency-changing of the incoming aerial signal—for further amplification at the intermediate frequency—introduces problems which are not usually encountered at ordinary broadcast frequencies. For example, circuits of more specialized nature are called for to amplify a weak signal at, say, 100 MHz and then to change its frequency to, say, 10 MHz, than those that are used for amplifying signals in the l.w. or m.w. broadcast bands and for frequency-changing to a relatively low value i.f., typically 470 kHz.

THE NOISE FACTOR

With the transistor in one guise or other taking over from the valve, the front-end noise problem has been considerably alleviated in recent years, especially with the advancing use of low-noise bipolar transistors and, in particular, field effect transistors (f.e.t. for short). All contemporary f.m. receivers thus adopt transistors. The f.e.t. is found in the more expensive hi-fi tuner and tuner-amplifier, commonly serving as the r.f. amplifier and sometimes also as the mixer, with a bipolar transistor as the local oscillator. We shall be dealing with transistor circuits later in this chapter, but for the sake of continuity and also because there are still many thousands of receivers and tuners working well with valves, many of which will require servicing for some years to come, apart from the academic interest, we really must start with the valve.

One of the chief problems involved is in securing a reasonable degree of unwanted noise. Noise in this connexion is not meant to mean disturbances picked up together with the signal due to atmospherics and interfering electrical appliances, but the noise contributed by the valves and associated circuits.

At Band II frequencies quite a lot of noise can be created by the r.f. amplifier valve following the aerial circuits. In the output circuit of the r.f. valve, therefore, there may be developed the amplified signal voltage and also a spurious voltage as the result of this valve noise. When the incoming signal is very weak it often happens that the noise voltage is equal to, or rises above, the amplified signal, and a limit is thereby imposed on the maximum useful sensitivity of the r.f. circuits.

Although this noise-voltage amplitude modulates the carrier of the signal and is not passed at full force through the receiver—as it would be if the receiver were of a.m. mode—it is nevertheless desirable to keep it at the lowest possible level. As a means of aiding in this respect, a triode valve is invariably used as the r.f. amplifier in f.m. receivers, as opposed to the more conventional pentode stage of a.m. receivers.

As the signal frequency is raised into the v.h.f. spectrum the noise output of a pentode valve increases, partly as the result of the electron stream within the valve dividing between the screen grid and the anode in a somewhat random fashion. This is known as the *partition effect*. The total noise output of the stage is also contributed to by other factors, such as the irregular flow of electrons between the cathode and anode of the valve, and also the irregular flow of electrons in resistors and wires associated with the circuit.

In fact, noise is produced in all components and in all stages, but provided the first stage possesses a relatively low noise factor, the signal is amplified above the noise, so the noise produced in the following circuits and valves is negligible in comparison and can thus be ignored.

Owing to the absence of the screen grid, a triode valve—in a specially arranged circuit—can often be used to greater benefit than a pentode at Band II frequencies. The triode is usually arranged so that its grid is earthed (connected to chassis) and the signal is applied to the cathode circuit. This means that the amplified signal in the anode circuit is in the same phase as the signal in the cathode circuit.

Apart from having the effect of lowering the input impedance of the stage, the earthed grid acts as an electrostatic screen between cathode and anode and thus breaks up the internal self-capacitance between these electrodes which, added to stray capacitances of the circuit and wiring, would otherwise cause the valve to act as an oscillator. This would be the case if the valve were used in the more conventional mode, with the cathode earthed and the signal applied to the grid.

Although the earthed-grid arrangement may not provide as much stage gain (amplification) as the earthed-cathode arrangement (owing to the production of a large measure of negative feedback—which occurs because the valve current flows through the input circuit) its low noise factor at v.h.f., coupled with its relatively low input impedance and ease of design, makes it eminently suitable for the r.f. stage of f.m. receivers.

The basic circuit of the earthed-grid r.f. amplifier is shown in Fig. 4.1. The aerial signal, from a dipole via a special feeder, is applied across the small coupling coil L1. The signal is thus induced in the cathode coil L2 which is resonated by C1. Resistor R1 provides the valve with cathode bias. The amplified signal appears across L3 which is tuned by C3. Variable tuning over Band II

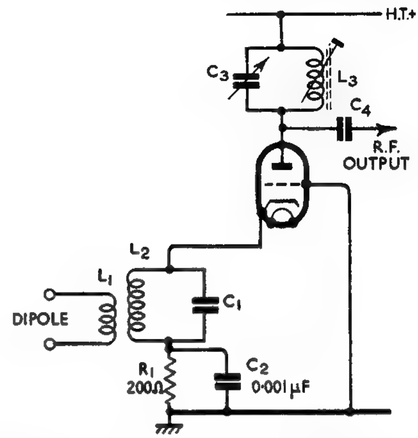


FIG. 4.1. *Basic circuit of the earthed-grid r.f. amplifier. The aerial signal is applied accross the coil L1.*

is performed either by varying the capacitance of C3 or by varying the inductance of L3. The signal is conveyed to the frequency-changer stage by way of the coupling capacitor C4.

Because of the low input impedance at the cathode circuit, the tuned-circuit L2, C1 is fairly heavily damped and consequently exhibits a wide response which embraces the whole of Band II, thereby obviating the need for variable tuning. The stage is tuned in the anode circuit since this is of considerably higher impedance.

The ratio of impedance between L1 and L2 permits correct matching of the feeder to the valve cathode circuit. Sometimes the cathode circuit is totally untuned (aperiodic), in which case the stage is arranged to match directly into the aerial feeder—the cathode input circuit can, in fact, be regarded as a step-up impedance transformer. In some circuits the cathode of the valve is tapped down L2 to avoid the excessive damping effect across the tuned circuit. When this is done a better input response is secured, but then it may be necessary to employ variable tuning to prevent fall-off of amplification towards the upper and lower limits of the f.m. band.

THE NEUTRALIZED TRIODE

It is possible to neutralize the grid/anode capacitance of a triode valve and employ it as an r.f. amplifier with the signal applied at the grid circuit without fear of the stage turning into an oscillator. This arrangement (Fig. 4.2) is sometimes used in favour of the earthed-grid circuit.

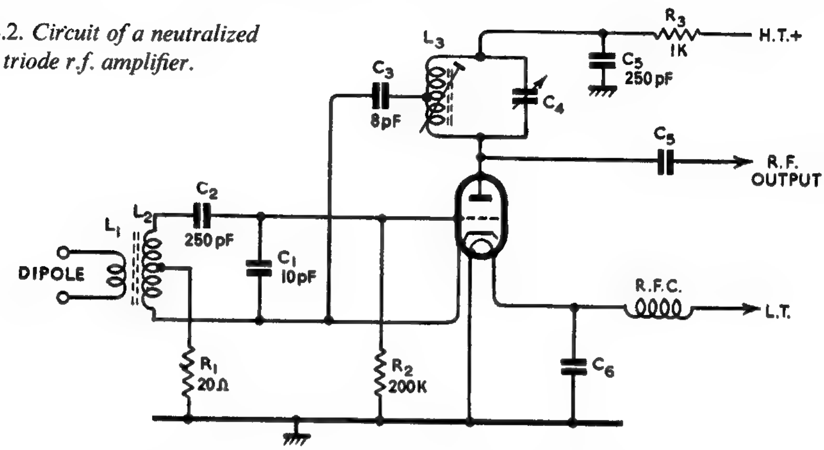
As will be seen from the circuit, there are a few extra complications, but these are sometimes thought worth while considering that the neutralized triode usually provides greater gain since it does not produce degenerative feedback.

As before, the signal brought in by the dipole is coupled to L2 via L1, but L2 this time is connected across grid and cathode. The cathode is virtually at chassis

potential by reason of the 20-ohm resistor, R1, which gives a certain amount of fixed bias to the valve. Capacitors C2 and C1 serve as coupling and for fixed tuning respectively, while R2 is a grid-return resistor.

The stage is neutralized by the inclusion of C3. This is required because a part of the amplified voltage in the anode circuit, due to the action of the anode-to-grid capacitance, is carried back to the grid circuit, after undergoing a phase

FIG. 4.2. *Circuit of a neutralized triode r.f. amplifier.*



change, and thereby produces conditions for oscillation (positive feedback); C3 counters this effect by feeding back into the grid circuit just the right amount of signal voltage from the anode, but in opposite phase, to cancel out the unwanted feedback. Clearly, C3 is of critical value, and if it becomes necessary to replace this capacitor during a servicing operation, it is essential to use a replacement component of exactly the same value which should be wired into the circuit in the same position and by the same lengths of wire.

Again, the anode circuit can be tuned by varying the values of L3 or C4. The grid circuit is very flatly tuned so that its response embraces the whole of the f.m. band without the need for a variable element. Capacitor C5 and resistor R2 decouple the anode circuit from the r.f. aspect, while C5 performs as the output feed capacitor. Sometimes it is considered desirable to filter the heater circuit of the r.f. valve—to obviate the possibility of instability—by an r.f. choke (r.f.c.) and capacitor (C6).

In practice only a small degree of amplification is given to the signal at v.h.f., irrespective of the type of r.f. stage employed. The main amplification takes place in the i.f. stages, after the frequency of the signal has been reduced by the frequency changer. Essentially, the r.f. stage of f.m. receivers serves as a buffer between the aerial and the frequency changer, and also allows accurate matching between the aerial feeder and the input circuit.

The securing of a good impedance match between the r.f. amplifier and the feeder ensures that the very maximum of signal is utilized. This is most important in districts on the edge of the service area where the aerial signal due to an f.m. station may be considerably below that due to the local l.w. or m.w. station.

Optimum signal transfer also enhances the signal-to-noise ratio of the stage

and thus permits application of the strongest possible noise-free signal to the frequency changer. This in itself also aids in maintaining a good overall noise factor, since the frequency changer can operate efficiently only when it receives a signal of substantial strength.

USE OF A PENTODE

Apart from the low-noise feature of an r.f. triode, this valve is also favoured by designers of commercial receivers because it can form one section of a double-triode valve, leaving the other section available for use as a frequency changer. A single valve, therefore, can be arranged to perform as an r.f. amplifier and frequency changer. Commercially, the saving of one valve, without detracting from performance, is economically desirable.

The constructor, though, is rarely faced with the same problems as the manufacturer, so in designs for the home constructor the r.f. stage often features a pentode valve in a fairly conventional circuit (Fig. 4.3). Here the aerial signal is

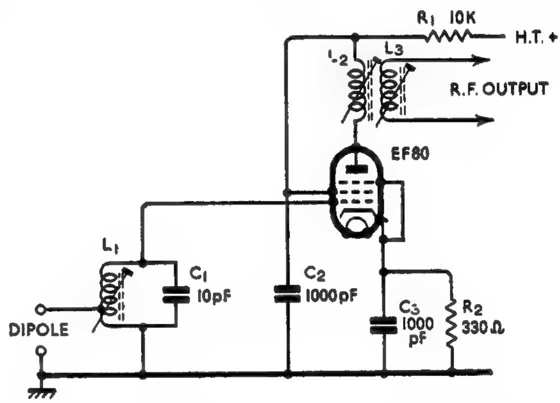


FIG. 4.3. *The r.f. stage sometimes features a high-slope r.f. pentode in designs for the home constructor.*

applied at the correct impedance point on the aerial coil L1. This coil is flatly tuned by the input capacitance of the valve and C1. The valve is usually of the high-slope (mutual conductance) variety, such as a Mullard EF80. This ensures that maximum stage gain is achieved in spite of the relatively heavy damping of the tuned circuits which occurs at v.h.f.

The amplified signal is developed across the r.f. coupling transformer L2, L3. This transformer is resonated by the self-capacitances of the windings, the stray circuit capacitances and the valve capacitance. The tuning of this circuit can be varied, within limits, by iron-dust cores which can be screwed in and out of the coil formers thereby providing a variable control of inductance. The resistor R2 gives the valve bias and C3 decouples the cathode at r.f.; R1 and C2 decouple the anode and screen-grid circuits.

A well designed pentode r.f. stage will not introduce excessive noise and will give a gain in excess of that of a simple triode circuit.

DOUBLE-TRIODE PUSH-PULL R.F. AMPLIFIER

The use of a double-triode in a push-pull earthed-grid circuit is sometimes featured in a design for the home constructor. Such a circuit is shown in Fig. 4.4.

The valve used is the low-noise double-triode, ECC85 being the Mullard designation.

The aerial feeder is isolated by the two capacitors C1 and C2, and the signal is applied across the two cathodes. The input circuit is balanced by the centre-tapped radio-frequency choke (r.f.c.) and resistors R1 and R2, which also

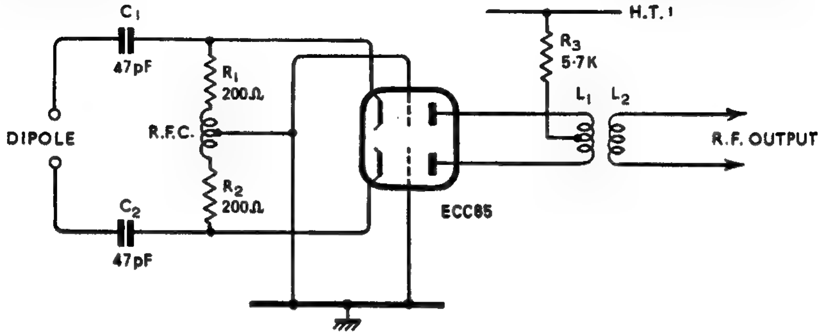


FIG. 4.4. *A double-triode in a push-pull earthed-grid circuit is also favoured by the home constructor.*

provide the valve with bias. The output circuit is similarly balanced by the centre-tap on L1 across which the amplified signal is developed. The circuit is quite straightforward and if well made should show an improvement over a single triode stage.

THE CASCODE CIRCUIT

The cascode amplifier is invariably featured in early two-band television receivers in view of its relatively high r.f. gain and low noise performance at v.h.f. It was decided to mention the circuit here since it functions as the r.f. amplifier on f.m., as well as on television in combined television/f.m. receivers.

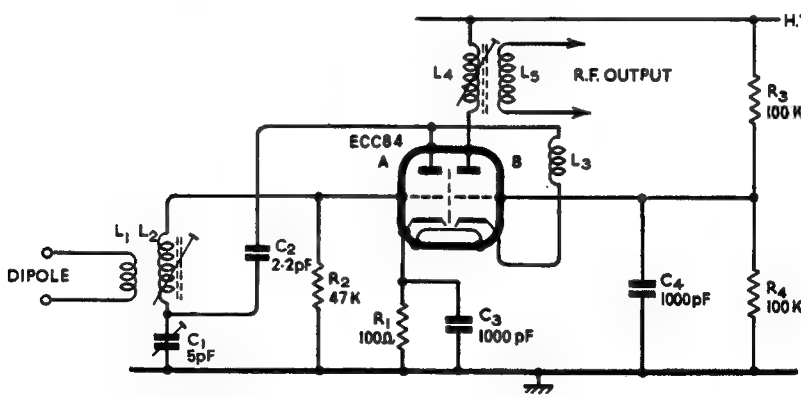


FIG. 4.5. *The cascode arrangement which features a neutralized triode in series with an earthed-grid triode.*

As will be seen from the circuit at Fig. 4.5, a double-triode valve is used; the first section (A) is connected as a neutralized triode and the second section (B) is connected in the earthed-grid mode. The triodes are thus treated as a single r.f. amplifier stage, and between them they are able to provide the low noise level

of a triode coupled with the relatively high gain and stability of an r.f. pentode.

The two sections are well screened from each other and both possess a high slope even when used with limited h.t. voltage. This latter factor is considerably advantageous, because it will be seen from Fig. 4.5 that the valve is best connected in series across a common h.t. supply.

The d.c. circuit between the two sections is from chassis (h.t. negative) through R1 to the cathode of section A; from the anode of section A through L3 to the cathode of section B; and from the anode of section B through L4 to h.t. positive.

Section A is biased by the volts drop across the cathode resistor R1. As the cathode of section B is held at a potential above chassis it is necessary to arrange the grid of the same section to be slightly less positive than the cathode so as to obtain a grid potential slightly negative with respect to cathode. One way of achieving this, as shown, is by connecting the grid to a tapping on a potential divider which is connected across the h.t. supply; resistors R3 and R4 perform this function.

From the r.f. viewpoint, the grid of section B is earthed (connected to chassis) by way of C4. The first section is neutralized by a capacitive bridge circuit comprising C1 and C2. Optimum noise performance is secured only when the stage is properly neutralized, and to this end C1 is often in the form of a small trimmer.

Stage gain is rendered relatively constant over the working frequency range of the circuit by the introduction of a peaking coil L3. This is usually wound with a few turns of heavy-gauge wire and supported in the circuit. With the coming of Band III television, a range of double-triode valves developed particularly for cascode circuits was introduced. A 6·3-volt version is the ECC84.

FREQUENCY CHANGER (V.H.F.)

In commercial f.m. receivers the second section of a double-triode valve is nearly always used as a self-oscillating frequency changer. This method of employing a twin valve lends itself to the production of a small r.f./frequency-changer tuning unit; this is desirable at very high frequencies since then the signal leads between the sections can be kept very short, and stray capacitance and inductance effects can be kept at the minimum.

As in the r.f. stage, a triode frequency changer is inherently less noisy than a pentode, hexode and heptode, though the conversion gain may be below that given by more conventional frequency-changer valves. Even so, at v.h.f. high conversion gains are not possible so there is little point in using a complex frequency-changer valve when a simple triode will do nearly as well, remembering still, that most of the amplification is given at intermediate frequency.

Fig. 4.6 shows the circuit of a typical self-oscillating triode frequency changer. The term *frequency changer* means that the valve performs the dual functions of local oscillator and mixer.

The tuned oscillator circuit comprises L1, C3. Capacitors C1 and C2 in series —giving a total of 3 pF—also add to the capacitance of the tuned circuit, being in parallel with C3. It will be seen that the r.f. signal is applied to their junction; the reason for this will be realized later.

The winding L2 is inductively coupled to L1 and forms the feedback or

reaction winding. Capacitor C5 is also in the feedback loop, from anode via L2 and L1 to grid. Both the f.m. signal and the oscillator signal appear at the grid of the valve. Developed in the anode circuit, across the i.f. transformer L3,

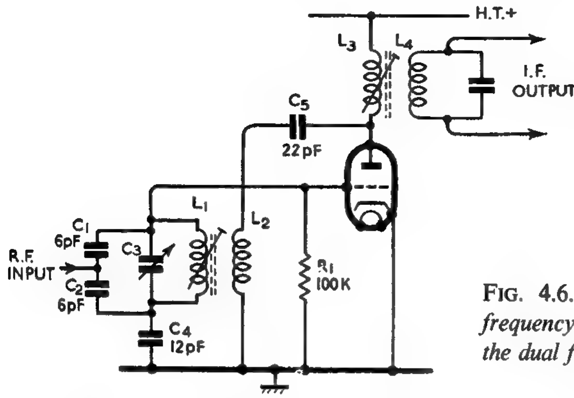


FIG. 4.6. *A self-oscillating triode frequency changer. The valve serves the dual functions of local oscillator and mixer.*

L4, is a frequency equal either to the signal frequency minus the oscillator frequency or to the oscillator frequency minus the signal frequency.

It is, however, general practice to arrange for the oscillator to work at a frequency equal to the signal frequency *plus* the intermediate frequency. The i.f. is thus equal to the oscillator frequency minus the signal frequency. This means that for a 10 MHz i.f. and a signal frequency of, say, 90 MHz the oscillator runs at 100 MHz.

PRECAUTIONS AGAINST OSCILLATOR RADIATION

With a receiver so adjusted, the second harmonic of the oscillator (200 MHz) falls in the middle of Band III, within the vision pass-band of Channel 10. It is essential, therefore, to ensure that all reasonable precautions are taken against oscillator radiation, which is readily propagated over relatively large distances at v.h.f., even at low power.

Direct radiation from the oscillator circuit is avoided by completely enclosing the tuning unit in a metal box and screening the frequency-changer valve. It is most important to ensure that all the associated screens and shields are securely repositioned after any servicing operation.

Back coupling from the oscillator circuit to the aerial is prevented by coupling the r.f. section to the frequency changer at the point of minimum oscillator potential. In the circuit at Fig. 4.6, the r.f. signal is applied to the frequency-changer grid circuit at the junction of C1, C2. When the circuit is properly balanced, this point represents the exact null point of the oscillator—where the oscillator potential is virtually at zero.

This is brought about because the oscillator circuit, including L1, L2, C1, C2, C3, C4 and the grid-to-anode capacitance of the triode, is designed as a balanced bridge so far as r.f. is concerned. Since one arm of the bridge is made up of the capacitance of C4, this component is often made variable in the form of a small trimmer, so that adjustment can be made for optimum balance of the oscillatory bridge circuit. Under this condition the junction of C1, C2 is

theoretically dead. Actually, due to unwanted—though unavoidable—couplings a little oscillator potential exists at this point, though it is usually of insufficient magnitude to cause any trouble.

When C4 is variable, it is accurately set at the factory and readjustment is rarely necessary except, perhaps, when certain key components in the oscillator circuit have been replaced. Adjustment should be carefully made for minimum oscillator signal at the receiver aerial terminals. Another way is by connecting a micro-ammeter in series with the oscillator grid-leak, R1, to measure the oscillator grid current, and then periodically shorting the junction of C1, C2 to chassis while noting the effect on the meter; C4 is in correct adjustment when no (or the smallest) difference in meter indication occurs between the shorted and open-circuit conditions. When making this adjustment, the meter should be inserted at the earthy end of R1 and shunted with a 1,000 pF ceramic capacitor.

In receivers where C4 is fixed, as in Fig. 4.6, an alteration in capacitance may not be noticed in the form of deterioration of receiver performance, but might well be brought to light by a complaint of excessive Band III television interference on a nearby receiver. If C4 deviated considerably from its normal value, then there is a possibility that it will show on the receiver as impaired tracking over the band (that is, the stations will fail to tune on the correct points marked on the scale) and low volume of certain stations, particularly when the receiver is operated in the fringe area.

TWO-STAGE FREQUENCY CHANGER

Nearly all v.h.f. frequency changers employ additive, or non-linear, mixing as distinct from electronic mixing which occurs in the electron stream of the valve. The electronic-mixing principle is adopted, for example, in the triode-hexode frequency-changer valve; here the r.f. signal is applied to the hexode control grid while the oscillator signal is applied to the oscillator injector grid, this being internally connected to the control grid of the triode oscillator. Because the hexode control grid and the oscillator injector grid are separated by the screen grid—which is virtually at chassis potential from the r.f. aspect since it is decoupled—neither grid has an undue influence on the other and mixing occurs in the electron stream flowing through the hexode section.

With additive mixing, both the r.f. signal and the oscillator signal appear on the control grid of the mixer section. This is illustrated in Fig. 4.7 which shows a two-stage frequency changer comprising a pentode mixer V2 and a separate oscillator V1. It will be seen that the r.f. signal, from the r.f. amplifier stage, is applied to the control grid of V2 through C7, and the oscillator signal is applied to the same electrode through C4.

With all systems of this nature the application of the oscillator signal causes grid current to flow in the pentode mixer section and thus results in non-linear amplification of the r.f. signal. The mixer section is a form of detector and the application of the two signals, when operating under non-linear conditions, results in the sum and difference frequencies of the r.f. and oscillator signals being generated in the anode circuit of the valve, as already considered.

Full efficiency in circuits of this kind is realized only when the mixer control grid receives the correct ratio of r.f. signal-to-oscillator volts. For this reason it

is important when making component changes to use the values stipulated by the designer.

Sometimes, instead of capacitive coupling of the oscillator signal, inductive coupling is favoured. Both arrangements are used in commercial receivers and in designs for the home constructor.

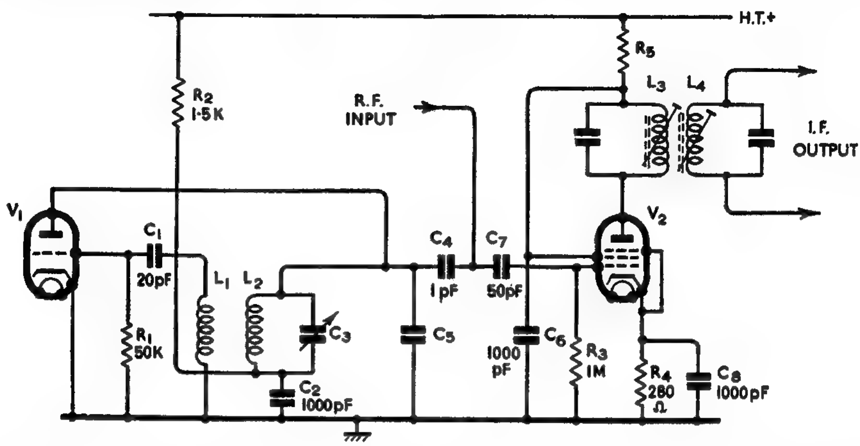


FIG. 4.7. *A v.h.f. frequency-changer circuit in which V1 operates as the local oscillator and V2 as the mixer.*

Although the circuit shown at Fig. 4.7 is representative of the majority of two-stage frequency changers, small details will be found to differ between designs. The local oscillator is usually of the balanced mode with respect to the r.f. coupling so as to prevent the oscillator signal from reaching the aerial. For reasons of economy the self-oscillating triode frequency changer is chiefly found in commercial receivers; the two-stage system is often featured in designs for the home constructor.

USE OF TRANSISTORS

Having now seen the problems associated with valves and the way in which they are arranged in r.f. amplifier and frequency changer stages, we are in a good position to examine similar circuits using transistors. It is not intended in this book to delve into the theory of transistors and associated semiconductor devices. Indeed, there just would not be sufficient room. Readers who are keen to secure a practical grounding in such matters, especially from the aspect of servicing, might find some of my other books useful, including *Radio and Audio Servicing Handbook* and *Hi-Fi and Tape Recorder Handbook*. Both of these have been completely revised to embrace transistors. There is also my *Rapid Servicing of Transistor Equipment*, dealing with transistors exclusively. All these books are published by Newnes-Butterworth, the publishers of this f.m. book.

A fairly simple f.m. front-end featuring three ordinary bipolar transistors (npn types) is shown in Fig. 4.8. TR1 is the r.f. amplifier, TR2 the mixer and TR3 the local oscillator. The r.f. amplifier is wired in common-emitter mode (similar to the "earthed cathode" of a valve amplifier), the aerial signal thus being applied to the base, via the coupling transformer T1 and the 5·6 pF capacitor.

The amplified signal is developed across the tuned circuit in which L1 is a part in the collector, and from there is fed to the base of the mixer TR2. TR3 local oscillator signal is also fed to the base of the mixer, via the 15 pF capacitor. The mixer operates in the common-emitter mode—the i.f. signal being developed across T2 (the i.f. transformer) in the collector circuit—while the oscillator

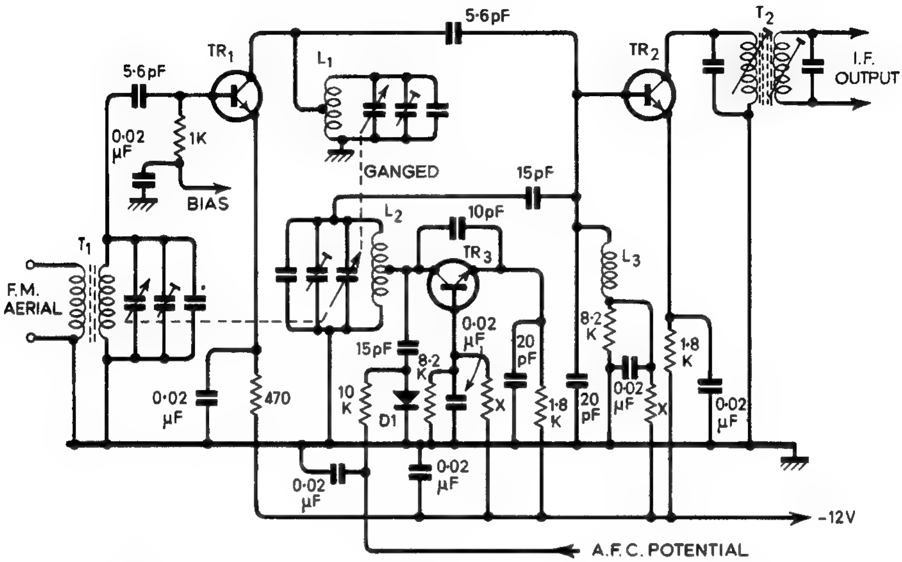


FIG. 4.8. *Transistorized f.m. front-end with a.f.c. It is noteworthy that facilities are often provided for switching off the a.f.c. when tuning a station so that the accurate tuning point can be established before the circuit comes under auto control. Without this it can be difficult to find the correct tuning point, depending on the efficiency of the system.*

itself is basically in the common-base mode (something like an "earthed-grid" valve circuit), the base being "earthed" signalwise by the 0·02 µF capacitor.

Oscillator and Mixer

A transistor in common-base mode exhibits a phase lag of about 90 deg. between the input and output currents at v.h.f., and it is partly for this reason that when the collector of TR3 is coupled to the emitter via the 10 pF capacitor oscillation occurs at a frequency determined by the tuned circuit, of which L2 is a part. This kind of oscillator circuit is used extensively in f.m. sets.

Heterodyne action occurs at the mixer as the result of the signal frequency and oscillator frequency signals being applied to the non-linearity existing between the base/emitter of TR2. The two signals combine to produce a complex waveform and T2 is tuned to select the i.f. difference frequency. Actually, when the waveform is applied across the base/emitter a part of it is suppressed by the non-linearity of the voltage/current characteristic, and it is this non-linear effect that results in the "mixing".

To secure the correct i.f. "difference" over the tuning range the variable capacitor tuning L2 is held exactly out-of-step with those tuning the input transformer T1 and the r.f. amplifier output inductor L1. The trimming capacitors

in parallel with the tuning gang sections make it possible to secure this condition at the top of Band II, while tracking at the bottom (low-frequency) end is sometimes achieved by adjusting the inductance of the coil and transformers. The local oscillator usually runs at a frequency equal to the i.f. plus the signal frequency.

Incidentally, L3 in the circuit is merely an r.f. "stopper", which "holds off" the oscillator signal from the base biasing resistors and circuits. Without this the oscillator signal would be severely damped by the biasing arrangement.

The base of the r.f. transistor, TR1, is biased by a potential derived from the i.f. stages, and this is often linked to the limited stage or stages so that TR1 is progressively turned down, so to speak, as the signal amplitude rises. This is a kind of a.g.c. linked to the limiting action which to some degree prevents TR1 from overloading, and running into non-linearity, in the presence of a strong aerial signal.

Automatic Frequency Correction

It will be seen that the oscillator tuning L2 etc. has in shunt with it diode D1 in series with a 15 pF capacitor. This circuit, in conjunction with the diode biasing feed, provides automatic frequency correction (sometimes called automatic frequency control). The diode, sometimes called a "capacitor-diode" or "variable capacitance diode", is the solid state equivalent of the variable reactance valve or ferrite modulator, mentioned in an earlier chapter. When such a diode (indeed, any junction diode, though the capacitor-diode is specially

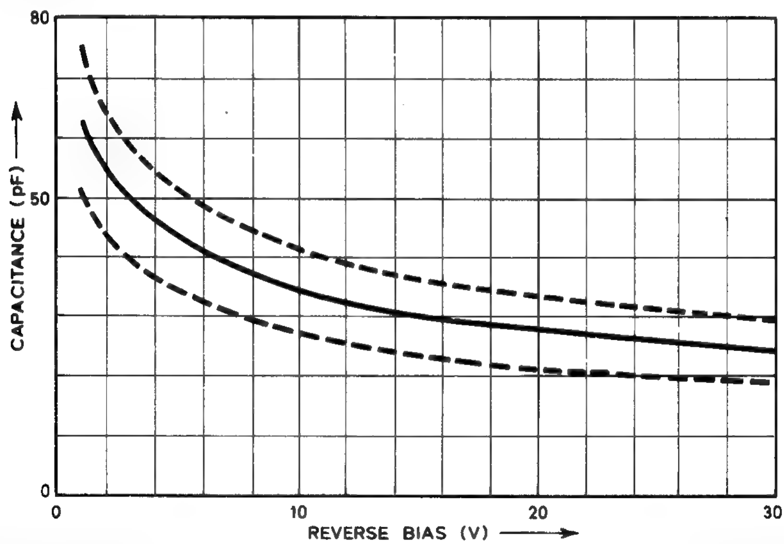


FIG. 4.9. *These curves show how the capacitance across the terminals of a capacitor-diode decreases as the reverse bias is increased. Full-line curve "typical" and broken-line curves revealing the possible "spreads", signifying maximum and minimum values.*

designed for the purpose) is biased in reverse it exhibits a value of capacitance across its terminals dependent on the degree of biasing. The greater the biasing, the smaller the capacitance, as shown in Fig. 4.9.

Clearly, therefore, with a capacitance like this in shunt with the oscillator coil L2 the oscillator frequency can be varied merely by changing the bias voltage. The diode is initially biased to produce the nominal oscillator frequency from a potential-divider network, but is also subjected to a bias obtained from the f.m. detector which, when the receiver is mistuned, can add to or subtract from the nominal bias, depending on the direction of mistuning. This f.m. detector bias, therefore, is highly suitable for automatic frequency correction, the action of which is to pull the receiver back into tune should it tend to drift off tune as the result of a drift in the frequency of the local oscillator for example.

The circuit in Fig. 4.10 shows a balanced ratio detector and from the discussion in Chapter 3 it will be recalled that when the signal circuits are exactly in

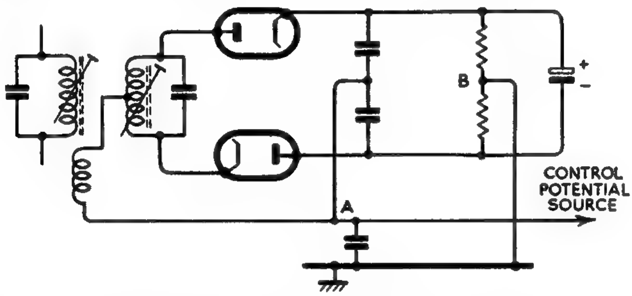


FIG. 4.10. *This kind of balanced ratio detector is found extensively in contemporary f.m. receivers, but using semiconductor diodes in place of the thermionic valve diodes shown here. It represents a convenient source of a.f.c. potential as the text explains.*

tune a state of perfect balance exists; that is, no voltage is present between points A and B. At frequencies above and below the correct tuned frequency, however, the voltage at A with respect to B (which, although shown at chassis potential here, is usually returned to the potential divider for applying a nominal reverse bias to the diode, as already mentioned) rises either positively or negatively, falling to zero when the i.f. signal at the ratio detector transformer is in sympathy with its tuning.

Whether the voltage at A rises positively or negatively with, say, an increase in frequency, depends on which way round the diodes are connected. If they are connected the wrong way round for the control (though this is unlikely to affect f.m. detection) the tuning error will be aggravated instead of corrected. In the circuit of Fig. 4.8 the oscillator frequency will increase (diode capacitance reduce) as the control potential swings less positive (e.g., in a negative direction).

It is noteworthy at this stage that capacitor-diodes are sometimes used to *tune* the r.f., mixer and local oscillator circuits instead of ganged capacitors. In this application the diodes are biased from a variable d.c. supply, such as from a potentiometer, this then serving as the tuning control.

Use of F.E.T.s

A very recent f.m. front-end, using f.e.t.s, is shown in Fig. 4.11. Here TR1 is an n-type-channel junction-gate f.e.t. arranged as the r.f. amplifier, driving into a mixer stage using the same sort of device, which is TR2. TR3 is the local

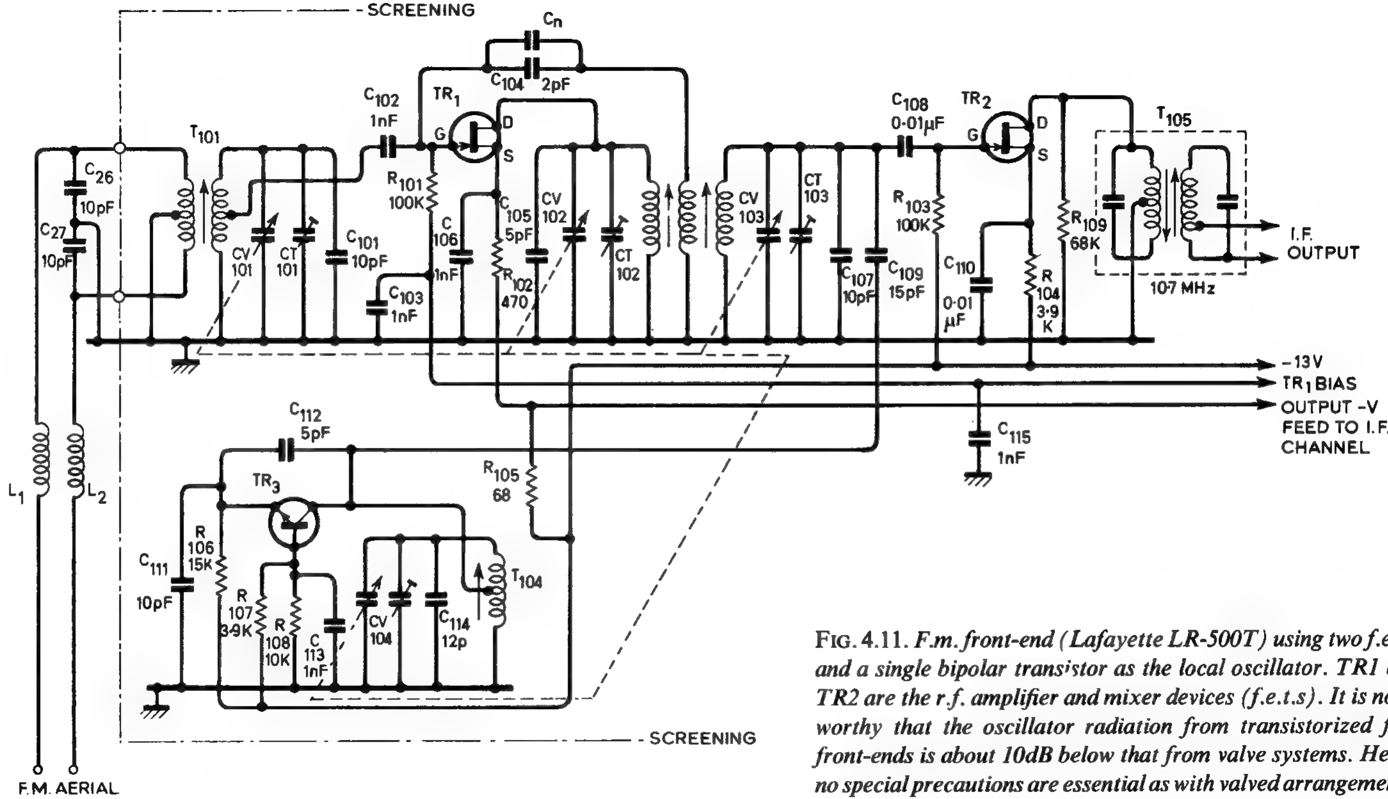


FIG. 4.11. *F.m. front-end (Lafayette LR-500T) using two f.e.t.s and a single bipolar transistor as the local oscillator. TR1 and TR2 are the r.f. amplifier and mixer devices (f.e.t.s). It is noteworthy that the oscillator radiation from transistorized f.m. front-ends is about 10dB below that from valve systems. Hence no special precautions are essential as with valved arrangements.*

oscillator, which utilizes an ordinary bipolar v.h.f. transistor wired in common-base configuration. The 5 pF oscillator feedback capacitor between the collector and emitter will be noted, and it will be appreciated that this stage is virtually indentical to that in the circuit at Fig. 4.8.

Local oscillator signal is applied to TR2 gate via a 15 pF capacitor, and to the same electrode is applied the signal frequency via the bandpass coupling transformer T102/103, from the r.f. stage. The mixer works in a manner similar to that of a bipolar transistor, the difference frequency being developed across the i.f. transformer T105.

Field-effect transistors have several desirable attributes over bipolars in v.h.f. front-ends. Firstly, they are less prone to overloading by strong signal, the early stages thus being kept reasonably clear of spurious signals which, under high signal conditions with bipolars, could arise from intermodulation effects. Secondly, they possess a high input impedance, since they are voltage-operated, like thermionic valves, instead of being current-operated like bipolars with the consequent low input impedance. Thirdly, taking into account the high input impedance, they are relatively less noisy than bipolars; and fourthly they are easier to gain-control than bipolar transistors. The various types of f.e.t.s are referred to in my companion volume, *Radio and Audio Servicing Handbook* (2nd Edition).

In other words, f.e.t.s are much more like thermionic valves in circuit and function than are bipolar transistors. Thus, the circuit in Fig. 4.11 is very similar indeed to a counterpart using thermionic valves.

The aerial signal fed to the primary of T101 via the filter comprising L1, L2, C26 and C27, giving the input circuit balanced 300-ohm characteristics, is tuned by the aerial section of the gang and tapped into the gate of TR1 via C102. The amplified signal is developed across the primary of T102 and is here tuned by a second section of the gang. Since an f.e.t. behaves rather like a triode valve under r.f. amplifier conditions, neutralizing feedback, to prevent oscillation or instability, is applied from a mutual winding on T102/103 back to the gate via the 2 pF capacitor and the small neutralizing capacitor marked C_n). The tuned and amplified signal is then fed to the gate of the mixer TR2 from across the secondary of T102, which is also tuned by a third section of the gang. The oscillator coil, T104, is tuned by the fourth section. The tappings down the transformer windings provide the correct impedance points for signal couplings.

Although this input is designed for 300-ohm balanced aerial feeds, it can be coupled easily enough to 75-ohm coaxial cable merely by disconnecting the input filter (balun) and coupling the inner coaxial conductor to one side of T101 primary and the braid to a chassis point. Under this condition only one half section of the primary winding is utilized, which increases the transformer turns ratio by two-to-one and the impedance ratio by four-to-one, thereby reducing the input impedance from 300 ohms balanced to 75 ohms unbalanced. However, many authorities prefer to use a low-loss balun to couple 75-ohm coaxial to 300-ohm balanced circuits—I do for one, especially when it is required to obtain the best possible signal transfer with the least noise (e.g., best signal/noise performance). The design for a suitable balun is given in my companion volume, *The Practical Aerial Handbook*.

Again, the r.f. amplifier is often biased from a potential derived from a limiter in the i.f. channel (this may be from an integrated circuit linear amplifier), as in Fig. 4.11. In this circuit the —13V input potential (relative to chassis which is connected to supply positive) is also fed out for connecting to the i.f. channel via R105.

Tuning Arrangements

A ganged variable capacitor is not uncommonly adopted for tuning the f.m. front-end stages, as in the circuits at Fig. 4.8 and Fig. 4.11. Quite a few tuner-amplifiers incorporate facilities for a.m. (medium-frequency) as well as for f.m.,

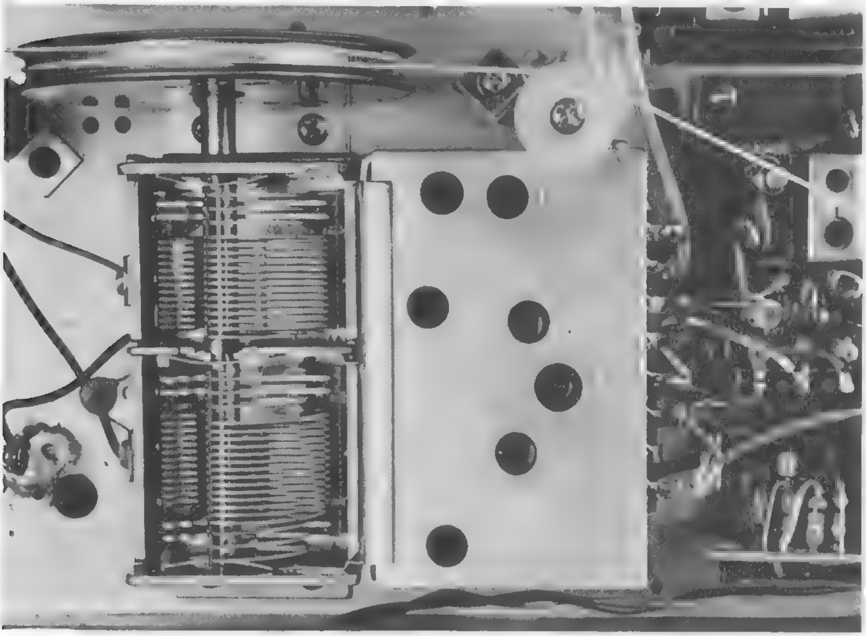


FIG. 4.12. *A.m./f.m. tuning gang. Here the two large capacitance sections tune the frequency changer and local oscillator of the a.m. section, while the three smaller sections tune the f.m. front-end.*

mono and stereo, and with these using capacitor tuning the gang is not uncommonly of a special type having, say, two sections for a.m. and three for f.m., as shown in Fig. 4.12.

Fig. 4.13 shows the permeability method of tuning, popular particularly in valved front-ends. Here a cord drive is secured to one end of the cores of the r.f. and oscillator coils, while the tension is maintained at the other end by means of coiled springs. The springs sustain a steady pull on the cores against the drive cord, which is passed over pulleys and wound round the shaft of the tuning control.

The illustrations in Fig. 4.14 relate to capacitor-diode tuning and they are all fully explained in the caption.

Some tuners and f.m. receivers, of course, still use the ordinary v.h.f. tuning gang, while others are based on switch tuning only, and where three or four

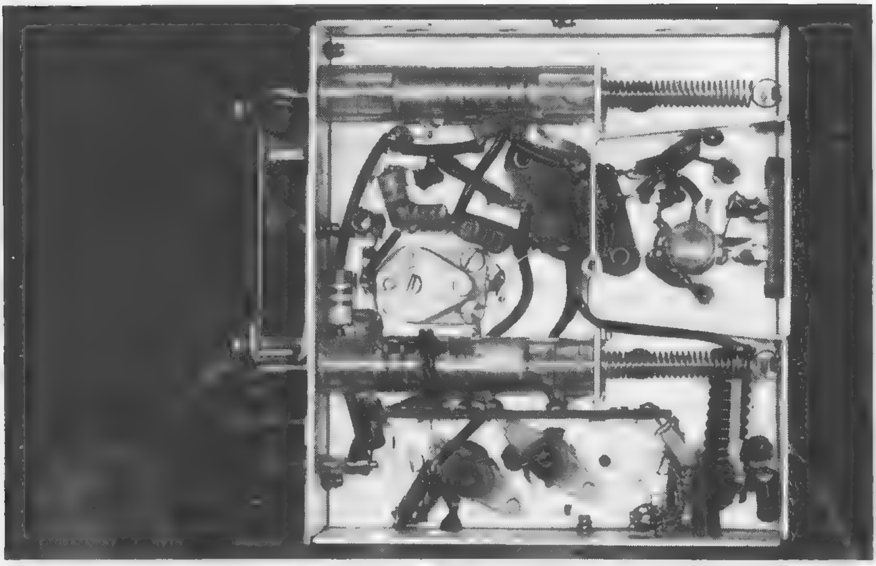


FIG. 4.13. *Inside view of f.m. front-end, showing the permeability system of tuning.*

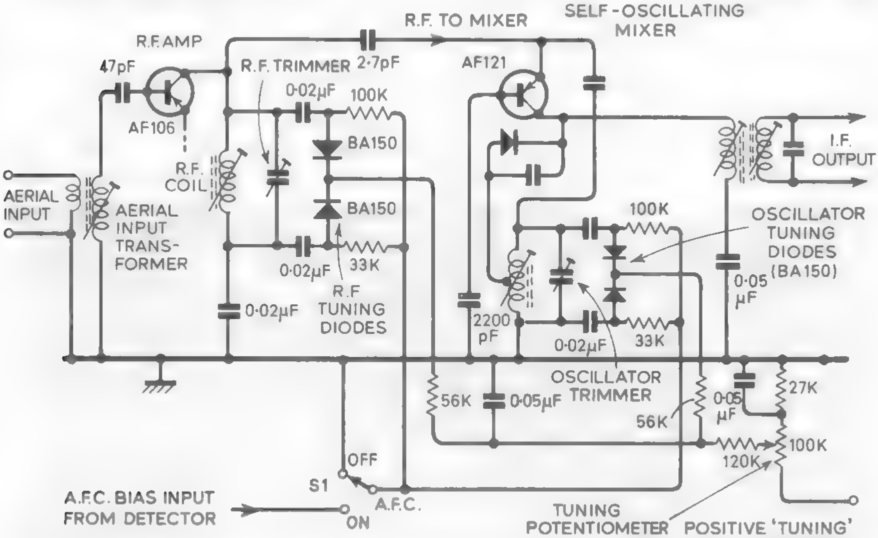


FIG. 4.14. *Capacitor-diode tuning. (a) Basic circuit showing r.f. and oscillator capacitor diodes. For the sake of clarity only the signal circuits are shown. Note the use of a* pair *of diodes (now a common practice) in each tuned circuit; these are reverse-biased by a positive input to their "cathodes", the tuning potentiometer providing variable bias and thus a tuning control. The diodes also operate the a.f.c., this being brought into action, or shorted out, by switch S1. This control potential, from the f.m. detector, is applied to the "anodes" of the diodes. Notice, too, in this circuit that the second transistor serves the dual functions of mixing and local oscillator. Some sets feature a third transistor for the local oscillator, as in Fig. 4.11 for example. The r.f. and oscillator circuits are tuned initially by dust-cores in the coils (for the low-frequency end) and by parallel trimmers (for the high-frequency end).*

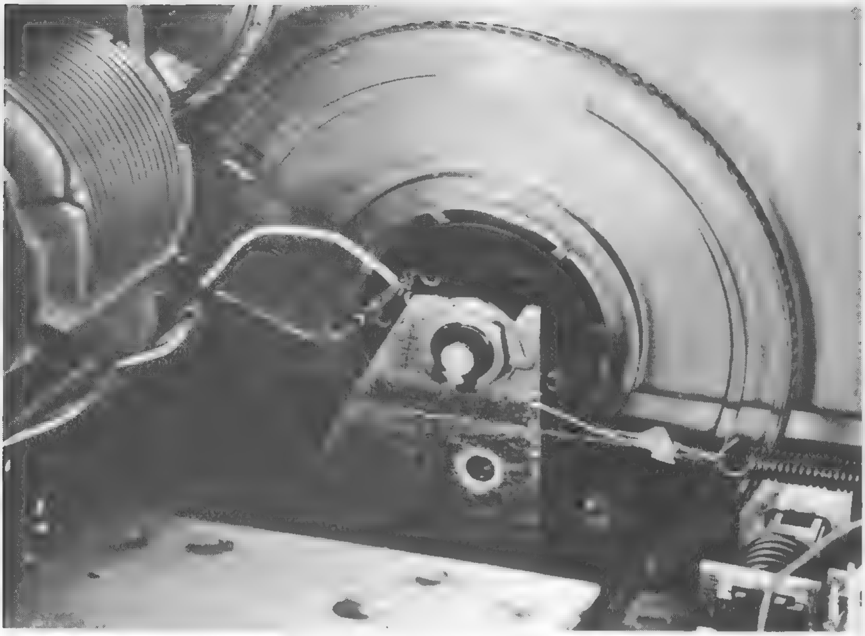


FIG. 4.14(b). *Capacitor diode tuning. Photograph of the "tuning potentiometer" in a set using capacitor-diodes. This is geared to the v.h.f. tuning drive and scale.*

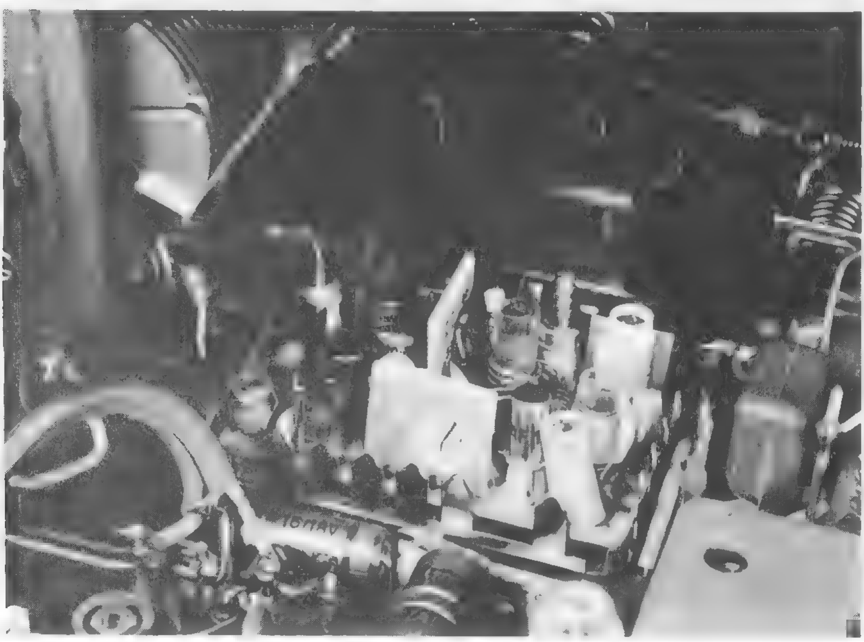


FIG. 4.14(c). *Capacitor diode tuning. Inside an f.m. front-end using capacitor-diode tuning.*

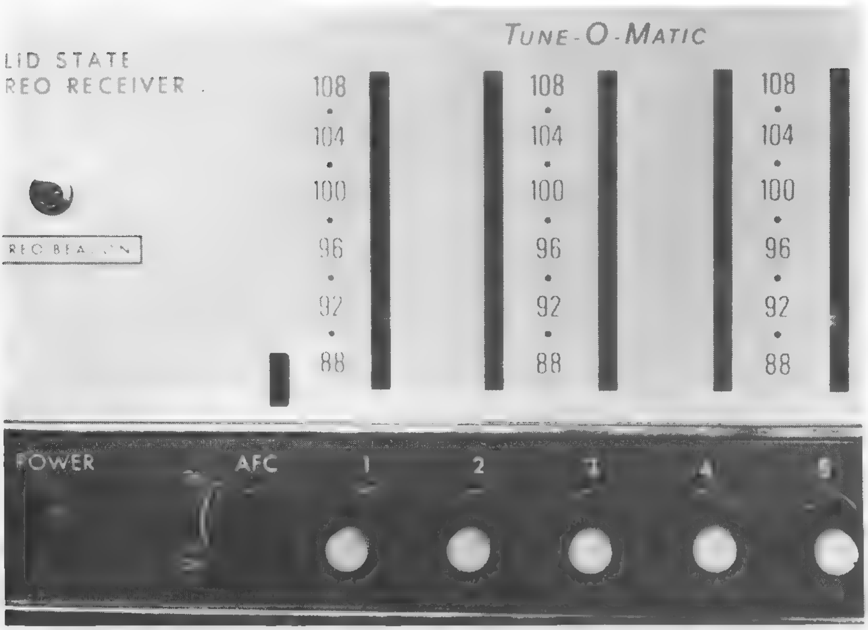


FIG. 4.14(d). *Capacitor diode tuning. Some receivers feature preset tuning for capacitor-diodes as this picture of the appropriate front section of the Fisher 160T shows. Here the required pretuned station is selected by pressing one of the white buttons. Pretuning is effected by pressing a button and then rotating the outer black part, an action which causes the related tuning indicator to traverse the scale which is calibrated in MHz from 88 to 108 (the whole of Band II). Notice, too, the a.f.c. on/off control and the "stereo beacon" or indicator as it is often called (see Chapter 7). Correct tuning to a station generally requires the a.f.c. to be switched off, after which the control can be switched on to prevent drift.*

stations of a local group are required the switch not uncommonly adjusts the oscillator tuning only by switching in or out increments of capacitance, thereby causing the tuning to shift over the 200 kHz f.m. channels. In this type of tuner the aerial and r.f. circuits are pretuned to the mean of the local group of channels.

CHAPTER 5

The I.F. Stages

TO OBTAIN the bandwidth required for the f.m. i.f. channel and to minimize second channel interference, a higher i.f. is demanded for f.m. than is appropriate for a.m. The standard f.m. i.f. is 10·7 MHz as against 470 kHz of the a.m. system.

We have already seen that the full benefits of the f.m. system are achieved in an i.f. bandwidth approaching 200 kHz, especially when the transmission is stereo-encoded (see Chapter 7), and that a limited bandwidth has the effect of attenuating the multiple sidebands of the signal carrying the higher modulation frequencies.

Owing to the extended f.m. bandwidth, it is not feasible to make use of the relatively low i.f. of 470 kHz to which we have become accustomed in a.m. receivers. To do so would require the i.f. stages to amplify linearly over a range of about 385 to 555 kHz. From the amplification point of view, this possibly could be arranged though the problems would be great, but apart from this, imagine the possibilities of interference from stations operating on frequencies within this pass-band should their signals gain admittance to the receiver's i.f. channel; this is a possibility.

So as to secure reasonable amplification over the desired f.m. pass-band, the ratio of i.f. to bandwidth must be kept relatively high. This ratio on a.m. receivers with a bandwidth of, say, 10 kHz and an i.f. of 470 kHz is about 47:1. To provide a ratio of the same degree with a pass-band approaching 200 kHz, a 10 MHz i.f., at least, is demanded.

IMAGE ACCEPTANCE

From the viewpoint of second-channel or image acceptance, a receiver tuned to a station at, say, 90 MHz would use an oscillator frequency of 90·5 MHz to give an i.f. of 500 kHz; this would mean that the second-channel would be at 91 MHz (the signal frequency plus twice the i.f.), which is well within the acceptance of the aerial and r.f. circuits. With a signal of the same frequency, but with an i.f. of 10 MHz, the second channel would be at 110 MHz; here a certain degree of attenuation would be offered to the unwanted signal by the pre-frequency-changer circuits.

The problem of second-channel or spurious signal pick-up of any kind is considerably eased on f.m. owing to the capture effect. Nevertheless, interference

of this kind may occur in fringe areas, so it is as well to design on the safe side. It must also be borne in mind that the three programmes (Radio 2, 3 and 4) come from a group of transmitters having the same location and power, and spaced exactly 2·2 MHz. This means that a receiver with a 1·1 MHz i.f. tuned to Radio 2 from Wrotham on 89·1 MHz would have a second-channel acceptance at 91·3 MHz, which represents, precisely, the frequency of Radio 3 from Wrotham.

Unlikely though it is for the choice of i.f. to be 1·1 MHz, the foregoing at least serves to illustrate the need for care in selecting the i.f., bearing in mind that there are at least three stations in a local group.

It has already been seen that low i.f.s are needed for pulse-counting circuits, but here the nominal i.f. is selected with extreme caution, and pre-set switch positions generally serve instead of the more conventional tuning dial.

The intermediate-frequency of 10·7 MHz was eventually decided after exhaustive consideration of the interference problem from all aspects. Consideration was given to the resulting oscillator frequency and its effects on other bands due to fundamental and harmonic radiation. Also taken into account were the signals in other bands and local channels which may combine with harmonics of the local oscillator and other spurious frequencies to produce an i.f. signal.

The frequency of 10·7 MHz is also sufficiently removed from 470 kHz to permit the a.m. and f.m. intermediate-frequency transformers to be connected in series in a.m./f.m. receivers without one adversely affecting the performance of the other.

PROBLEMS ASSOCIATED WITH A 10 MHz INTERMEDIATE FREQUENCY

The securing of sufficient amplification over the relatively wide pass-band economically, represents the chief problem relating to a 10 MHz i.f., for where adequate gain is yielded by a single i.f. stage at 470 kHz with a pass-band of some 10 kHz, two, or sometimes three, stages are called for at 10 MHz where linear amplification over 170 kHz is demanded.

These two factors, bandwidth and gain, are related in that stage gain becomes progressively more difficult to obtain as the nominal frequency is raised and when a wide pass-band is required. If, for example, a 50-kHz bandwidth *could* be employed then, even at 10 MHz, a single i.f. stage using high-Q transformers and little or no damping would probably be sufficient to provide all the necessary amplification. Moreover, with a.m. signals a reasonable degree of pre-i.f. amplification of the signal is feasible, but, as already seen, at v.h.f. we have to rely more on the amplification given at intermediate frequency.

The latest receivers and tuners use transistors and some integrated circuits (i.c.s), but before we get on to these let us investigate some of the valve circuits, of which many thousands are in use even in this transistor era.

In receivers which feature a Foster-Seeley discriminator, or a discriminator not endowed with the property of amplitude limiting, the amplification preceding the limiter must be large enough to provide a signal greater than the grid base of the limiter valve from even the weakest signal it is required to receive.

In some commercial receivers two i.f. stages are used, the final one being in the form of a limiter. Both stages usually employ valves with a fairly high mutual

conductance, but since the second stage also acts as a limiter it usually contributes only a little to the overall gain of the i.f. channel. If both i.f. stages were designed for maximum gain, difficulty would be encountered in maintaining complete stability of the i.f. channel and, in addition, it would be a problem to adjust the i.f. transformers to give the desired i.f. response curve. Moreover, the overall amplification might be too high for normal purposes, bearing in mind that f.m. receivers are not usually designed for long-distance reception but are arranged solely for the interference-free reception of local f.m. stations.

Since two valves working normally provide too much gain and a single valve too little gain, the second stage—even in receivers using a ratio detector—is often arranged as an amplitude limiter. This is a good thing since it doubly ensures rejection of amplitude disturbances.

In some home-constructor designs a single i.f. stage is often advocated where the receiver is to be used close to a powerful station, and provisions are made for the inclusion of an additional stage where the receiver will be used under fringe-area conditions.

To sum up, the degree of amplification necessary at i.f. depends to a large extent on the strength of the local f.m. signal; the design quality and economic structure of the receiver or adaptor represents another factor, as also does the type of discriminator used. Generally speaking, the ratio detector, being somewhat less efficient than the Foster-Seeley circuit, requires a greater signal for optimum operation than does the Foster-Seeley discriminator, but by taking into account the low amplification of the limiter stage, which is an essential embodiment in receivers using the Foster-Seeley circuit, the overall efficiency—so far as commercial receivers are concerned—is often greater when the ratio detector is used. Some elaborate designs feature two i.f. stages, arranged for relatively low-gain operation, plus a limiter which does not contribute at all to the overall amplification; it may in fact attenuate the signal.

From the normal design aspect, the choice of 10·7 MHz lends itself to relatively easy production of the i.f. channel. The usual impedance damping factors of the circuits and valves on a 10 MHz tuned circuit naturally provides a bandwidth of the order of 150 kHz without the necessity of stagger-tuning and the artifices which are familiar to the television engineer. This is brought out by the fact that reasonable operation of f.m. receivers can be expected simply by tuning the i.f. stages for maximum output, disregarding for the time being the discriminator transformer. If the same were done on the vision channel of a television set, one would be fortunate to secure a pass-band of 1 MHz, remembering that 5 MHz is the frequency usually stipulated here.

Nevertheless, the simple peaking of the i.f. transformers is not a good thing if full advantage is to be taken of the extended audo range of the system. The aim should be in obtaining a flat response curve over the i.f. pass-band so as to avoid cutting the higher audio frequencies; the peaking of transformers usually gives a peak to the response curve in the centre of the pass-band, and relative to this attenuates the a.f. signals contained in the sidebands representing the higher-order modulation. There is, too, the danger of producing a response curve which is non-linear and which possibly features a bulge at one side. It will be learnt later how the response curve can be viewed on the screen of a cathode-ray tube

and how this aids in securing the desired response curve without simply peaking the transformers for maximum output. Selectivity, too, is very important, especially for stereo.

TYPES OF COUPLING

Intermediate-frequency transformers for f.m. receivers are made much after the style of i.f. transformers for a.m. receivers. Apart from the smaller number of turns on the f.m. transformers, however, the coupling factor is somewhat more critical than is general with a.m. components.

Fig. 5.1 shows an i.f. amplifier stage, in basic form, of a typical f.m. receiver or adaptor. Here L1 and L2 represent the primary and secondary windings of

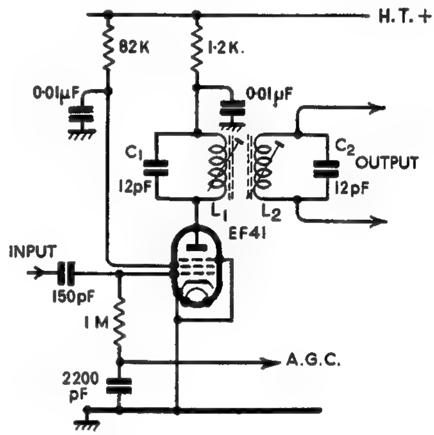


FIG. 5.1. *Circuit of i.f. amplifier stage of f.m. receiver adaptor.*

the i.f. transformer which are tuned in part by C1 and C2 respectively. These capacitors are usually contained in the can which serves to screen the transformer windings. The self-capacitance of the windings and the capacitances of the wiring and valves also contribute to resonate L1 and L2 to the correct frequency. Final tuning and balance of the circuits are achieved by adjusting the inductance of L1 and L2 by means of iron-dust cores which can be screwed in and out of the coil formers.

Now, the overall response curve given by L1 and L2 depends to a large extent on the degree of coupling between the two windings. If the coupling between them is loose—that is, if the two windings are sufficiently far apart so that the coupling is small—then, when the two circuits are tuned to the same frequency, the response curve will be a single peak, such as is shown in Fig. 5.2 (a). The broken-line curve gives a relative idea of the shape of the response curve due to a single tuned circuit.

When the coupling is increased, such as would occur by moving the windings closer together, the voltage induced across L2 increases until a point is reached when the voltage across L2 remains constant. Still gradually increasing the coupling causes the top of the response curve to flatten out as shown at Fig. 5.2 (b). If the coupling is further increased, the top of the response curve breaks into two peaks with a trough between, as shown at Fig. 5.2 (c). The diagrams thus illustrate three essential modes of coupling—loose, intermediate and tight.

An artifice used by designers is in mixing such couplings so as to secure an overall response curve which has a relatively flat top and steeply rising sides. How this may be done is depicted in Fig. 5.2 (d). The response of one transformer which has a loose-to-intermediate coupling is added to the response of

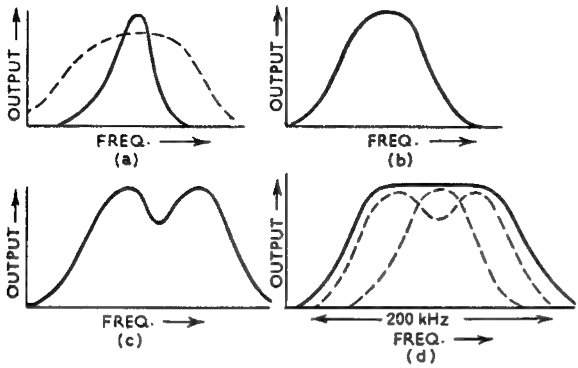


FIG. 5.2. *At (a) is shown a response curve due to an i.f. transformer with loose coupling. Intermediate and tight couplings are shown at (b) and (c); (d) shows the effect of adding two response curves due to different coupling factors.*

another transformer which has a tight coupling, so that the peak of one fills the trough of the other, giving an overall response curve of the required width, as shown by the full-line curve.

Fig. 5.3 gives a general idea of the construction of f.m. intermediate-frequency transformers. A common former is used to carry both primary and secondary

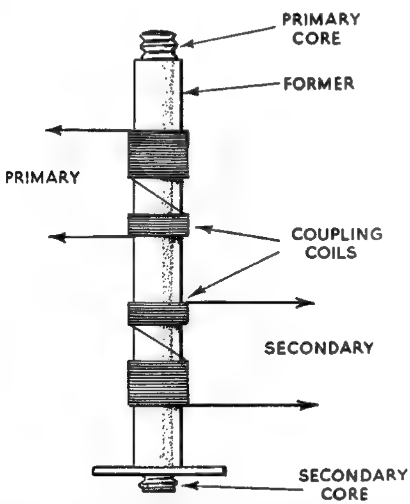


FIG. 5.3. *General construction of an i.f. transformer.*

windings; two dust cores are used, one at the top and the other at the bottom of the former. Since the primary and secondary circuits are tuned by adjusting the position of the cores in the former, their movement tends to affect the pre-arranged coupling factor.

As a means of alleviating this effect to some extent, each coil is sometimes separated to form two series-connected coils, as shown, so that the smaller section of each winding serves as a coupling coil outside the direct influence of the dust cores. The cores thus provide sufficient adjustment of inductance by traversing within the larger section of each winding.

This method makes the design of the transformers a little more difficult, but it is often considered worth while, and it was adopted in one Mullard f.m. tuner unit for the home constructor.

Full details for the construction of i.f. transformers are given in f.m. designs for the amateur, and in view of what has been said regarding the critical coupling factor it is as well to follow the details faithfully. It is outside the scope of a book of this nature to discuss principles of design to any large degree, but the following information may be helpful to home constructors. Roughly, 23 to 30 turns of 28 s.w.g. single-silk-and-enamelled copper wire close wound on an Aladdin type PP5937 former will tune over a range of 200 kHz at 10·7 MHz with a fixed tuning capacitor of about 50 pF.

The precise number of turns and the value for the capacitor depend on the magnitude of stray capacitances, due to the wiring and valves, reflected across the coil. As we have seen, the spacing between the two coils is tied up with the general design, though for an intermediate coupling factor the distance often ranges between 3/16 and 5/16 of an inch (5–8 mm).

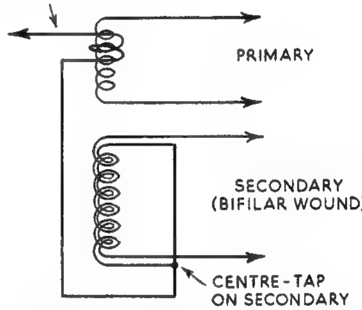


FIG. 5.4. *Illustrating the bifilar method of winding the secondary of the discriminator transformer. When the transformer is for use with a ratio detector, the tertiary winding is wound on top of the primary at the end remote from the secondary.*

The discriminator transformer is also a critically designed component whose coupling factor has to be adjusted with extreme caution. This is usually made on a common former as with the i.f. transformers and—when it is for use with a ratio detector—the tertiary winding is tightly coupled to the primary by being wound on top at the end remote from the secondary.

The precise value of coupling used, together with the number of turns for the tertiary winding, however, is again dependent on other circuit constants, which are selected before the coupling is finally set to the desired value. It is general practice to adopt the bifilar principle when winding the discriminator transformer secondary. The idea is to wind the secondary with two separate lengths of wire close together but insulated from each other, as shown in Fig. 5.4. The four ends are then sorted out so that the point where the two coil sections are connected in series forms the centre-tap of the secondary winding. This is connected to one end of the tertiary winding in accordance with the circuit of the ratio detector.

Briefly summarized, the design of the i.f. transformers, in relation to the i.f. amplifier stages, has to provide for (a) a pass-band wide enough to permit the undistorted passage of all significant sideband components of the f.m. signal, (b) sufficient selectivity so as to avoid the i.f. channel being responsive to signals

in an adjacent channel, (c) the minimum of distortion within the pass-band limits of the i.f. channel—that is, the amplification over the pass-band is required to be essentially linear without undue peaks and troughs, and (d) slightly wider bandwidth than is required by the signal so as to prevent excessive distortion as the result of detuning due to drift of the local oscillator.

With a.m. reception, insufficient bandwidth, or distortion of the response, results in attenuation of the higher audio frequencies. With f.m. reception, apart from this, considerable harmonic distortion is likely to be present in the reproduction, which might be considerably more disconcerting than a limited frequency response. This is one of the reasons why correct alignment of f.m. receivers is essential.

INTERMEDIATE-FREQUENCY AMPLIFIER STAGE

As revealed in Fig. 5.1, there exists very little difference between an a.m. and an f.m. intermediate-frequency stage from the circuit point of view. This should not deceive, however, for a big problem in the design of f.m. intermediate-frequency amplifiers lies in avoiding an undesirable amount of regeneration (positive feedback). The valves used in current designs have quite a high mutual conductance, and since these stages often use transformers with a Q factor approaching the maximum allowable by bandwidth considerations, the possibility of regeneration is considerably aggravated. In addition, it must be remembered that three valves may be used to amplify at i.f., as opposed to one or two in a.m. sets.

NEUTRALIZING THE INTERMEDIATE-FREQUENCY AMPLIFIER

Quite often regeneration occurs as the result of anode-to-grid capacitance in a single valve. Where this is the case, the neutralization of the valve responsible provides a simple solution without detracting from the overall performance.

There are a diversity of methods of achieving neutralization, but a simple method is shown in the circuit in Fig. 5.5. The capacitor shown in broken line

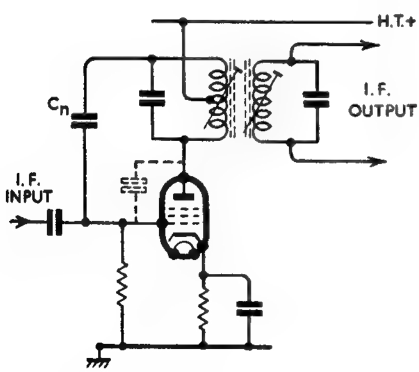


FIG. 5.5. *A method of neutralizing the i.f. amplifier. The voltage fed back from the anode circuit by way of the capacitor C_n cancels the offending voltage which reaches the grid via the grid/anode capacitance of the valve.*

represents the capacitance between the anode and grid of the valve, which may also be contributed to by stray circuit capacitances, and which is the primary cause of instability due to regeneration.

The idea of neutralization in the r.f. amplifier is in feeding back part of the output voltage at the anode in such a way that it appears in opposite phase to the offending regenerative voltage at the grid.

This is accomplished in Fig. 5.5 by applying h.t. for the anode of the valve at a tap on the primary of the i.f. transformer, and feeding back a correct portion of the signal—in anti-phase—to the grid circuit by way of the neutralizing capacitor C_n. In order to secure perfect cancellation of the offending regenerative voltage, the value for C_n has to be selected with extreme care. To aid in this respect, it is often a small pre-set trimmer which can be adjusted to counter the effects of phase shifts, stray capacitances and inductances which inevitably occur in i.f. amplifiers, which may not have exactly the same magnitudes between models of the same series, and which make accurate determination of the value for C_n extremely difficult.

Probably a more economical method of achieving neutralization, and one which avoids the necessity of tapping the primary of the i.f. transformer, is shown in Fig. 5.6. The capacitance distribution in the form of circuit strays and

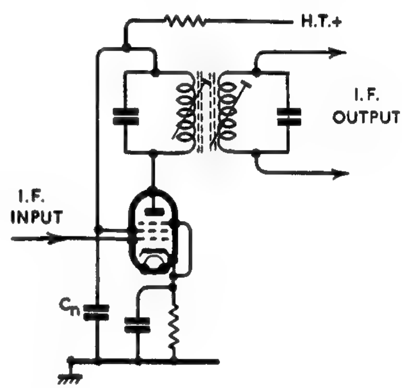


FIG. 5.6. *A less costly method of neutralization makes use of a capacitor, C_n, of critical value to form a balanced bridge in conjunction with the distributed capacitances of the amplifier circuit.*

those relating to the valve and capacitor C_n form a bridge circuit which can be balanced to counteract regeneration by judiciously selecting the value of C_n.

Although C_n, in this circuit, looks very much like an ordinary decoupling capacitor, it must not be treated as such during a servicing operation demanding its replacement. An average value for this component is 3,000 pF, but this value may be different in some designs.

The same reasoning applies to all capacitors in the i.f. amplifier, since most of them are chosen carefully to suit a 10 MHz i.f. With decoupling capacitors it is possible that their values are selected to resonate stray circuit inductances within the i.f. spectrum so as to achieve enhanced decoupling at the operating frequency. Even the wires on the capacitors may be cut to a critical length with this aim in mind.

Usually, either the r.f. amplifier valve or the i.f. amplifier valve (or both) is in receipt of an a.g.c. potential either from the discriminator or from the limiter stage. It has already been shown how the ratio detector provides an a.g.c. voltage, and it is shortly to be shown how the limiter can also provide such a voltage.

One method of applying the a.g.c. potential to a controlled i.f. amplifier valve is shown in Fig. 5.1. Here the signal is capacitively coupled to the control grid of the valve through the 150 pF capacitor. The 1-megohm resistor provides a grid return path through the a.g.c. system, and the 2,200 pF capacitor serves to decouple the a.g.c. line.

The Mullard EF41 and the EF85 valves both possess variable-mu characteristics and have a high mutual conductance making them eminently suitable for use in the controlled stage of an f.m. receiver or tuner unit.

AMPLITUDE LIMITERS

It is a basic requirement for all high-quality f.m. receivers and f.m. adaptors to provide good amplitude limiting, either wholly in the i.f. chain or party in the i.f. chain and partly in the detector. The ratio detector, properly designed and tuned, provides a large degree of amplitude limiting without requiring the assistance of any other section of the circuit. In some cases, the inherent limiting factor of the detector is considered sufficient; in other cases, it is supplemented by a limiter stage in the i.f. amplifier chain. With the Foster-Seeley circuit, one limiter stage at least is essential. Often two limiters in cascade are employed in larger more expensive receivers.

In Fig. 5.7 (a) and (b) is shown a typical amplitude-limiter circuit, such as that employed in the majority of domestic receivers. The circuit section at

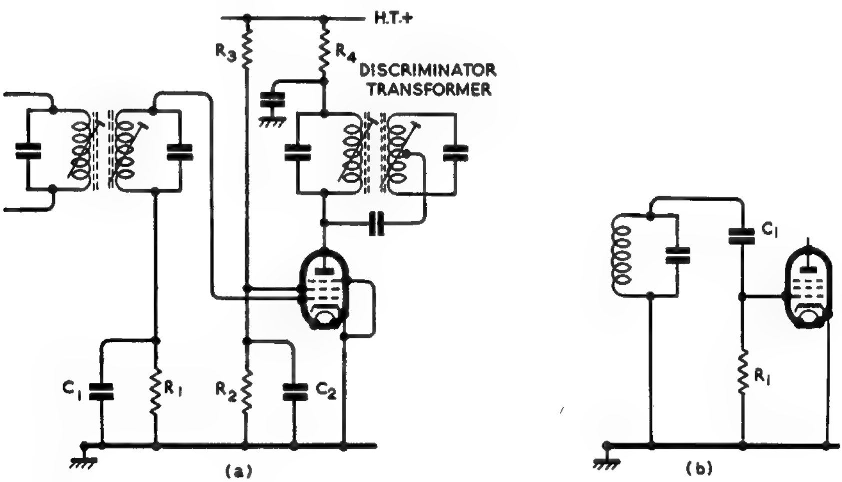


FIG. 5.7. *(a) An amplitude-limiter circuit in which C1 and R1 form the time-constant circuit. At (b) is shown an alternative method which results in greater i.f. transformer damping.*

(b) illustrates an alternative method of connecting the time-constant circuit C1, R1. The function of this circuit is the same in each case, but where the time-constant is introduced in the earthy end of the secondary of the i.f. transformer at (a), the damping of the transformer as the result of grid current is less than with the arrangement shown at (b). When two limiters are used, the first may be arranged as at (a) and the other as at (b).

A general idea of the working of this kind of limiter is given in Chapter 2. It will be remembered that the valve is given a short grid base by operating it with comparatively low anode and screen-grid potentials. To provide a stabilized screen voltage, the screen grid is fed from a potential divider formed of R2 and R3. Capacitor C2 serves to decouple the screen grid at i.f.

The values of R2 and R3, typically 30,000 and 100,000 ohms respectively, supply the screen grid with in the region of 50 volts, while a value of something like 50 to 100,000 ohms for R4 gives a similar voltage at the anode.

The time-constant circuit C1, R1 also contributes to the limiting action, and this is very much after the style of a leaky-grid detector. When the signal exceeds the grid base, which is usually set at 1 volt or so, grid current flows in R1, and C1 acquires a charge so that the grid is biased negatively. Now, if the signal increases, the average negative bias on the grid will also increase, thereby having the effect of holding the output signal constant. Moreover, because of the low values of anode and screen voltage, the voltage swing at the anode is severely restricted, and once the input signal exceeds the grid base of the valve there will be hardly any increase in output voltage at the anode.

When it is realized that at least 2 volts peak is required to cause satisfactory limiting in a typical receiver, one can appreciate the importance of having as much i.f. gain as possible prior to the limiter stage. The greater the input signal voltage at the limiter, the better the limiting, of course. For really efficient limiting, a signal in the region of 10 to 12 volts peak is required. This is not easily obtainable though, particularly in fringe areas where the signal given to the receiver might well be less than 200 microvolts (0·0002 volt). In such conditions the limiter rarely functions as such, for it is unlikely that the peak signal at the limiter exceeds the grid base of the valve. This may not matter much in relatively interference-free zones, but in heavily built-up areas, the resulting electrical and ignition interference might well be sufficient to detract from the whole aspect of the f.m. system.

It is possible to produce f.m. receivers having outstandingly high gains which would probably limit successfully with an aerial signal of a couple of microvolts or so, but these are usually confined to the purpose of communications where cost is less important than in the domestic case. Latest hi-fi sets limit at that level.

Better limiting in fringe areas can be obtained by using two limiters in cascade, but apart from the more expensive type of receiver or f.m. adaptor, the cost is, in general, against installing cascade circuits in the more popularly priced equipment. Where two limiters are used, however, care has to be taken over their coupling so as to preclude the possibility of introducing amplitude modulation. Ordinary transformers are sometimes used, while in other cases shunt capacitance coupling is favoured, where single tuned circuits are capacitively coupled.

The values for the time-constant circuit C1, R1 are arranged so that the grid bias on the limiter valve is able to follow the changes in amplitude of the input signal. This stipulation usually means that a time-constant in the region of 2 to 3 microseconds is most suitable. Thus, R1 may be, say, 50,000 ohms and C1 50 pF. A shorter time-constant may not produce a bias voltage large enough to control the amplification of the limiter valve in the face of amplitude fluctuations of the

input signal, while a longer time-constant may detract from the limiting given during unusually heavy bursts of car-ignition interference. A compromise time-constant is found in receivers to match a certain method of design and to provide optimum limiting under a diversity of reception conditions. It should be remembered that domestic receivers must be designed to operate with reasonable success not only in areas of high signal strength with little or no interference, but also in fringe areas where the magnitude of interference may approach that of the signal.

It must be stressed that no matter how well the limiter stage is designed, amplitude disturbances of the signal will not be totally suppressed unless the tuning of the i.f. transformers and discriminator transformer is accurately performed. The converted carrier, or nominal i.f., must fall in the centre of the i.f. pass-band and in the centre of the straight portion of the discriminator characteristic curve. Even then, when the receiver is tuned with utmost care, under adverse reception conditions bursts of heavy ignition interference may break through on the reproduction. When this happens in a heavily populated district in a fringe area one should not take the set to pieces to rebuilt it, but be content in knowing that the trouble is due to lack of signal coupled with a poor local signal-to-interference ratio.

The circuit in Fig. 5.8 illustrates how the time-constant resistor is made up of two series-connected resistors to form a potential divider from which an a.g.c.

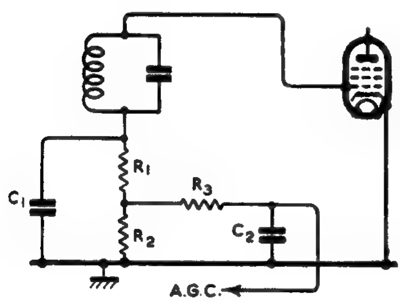


FIG. 5.8. *The time-constant resistor can be made up of the resistors R1, R2 to form a potential divider from which an a.g.c. voltage may be taken.*

voltage may be taken. The values for R1 and R2 are chosen so that the correct amount of bias voltage, due to the charge on C1, is available for application to the controlled valve or valves. Since the voltage across C1 follows the amplitude of the signal at the limiter, a signal fade or increase will result in a decrease or increase in negative a.g.c. voltage, and thus give rise to a corresponding increase or decrease in the overall gain of the receiver to counter the effect of the signal variations.

Resistor R3 and capacitor C2 serve to rid the a.g.c. voltage of undesirable pulses. These components also ensure that the control voltage is free from i.f. voltage which may result in regeneration.

TRANSISTOR I.F. STAGES

Fig. 5.9 shows a simple transistor f.m. i.f. channel in which TR1 is the first transistor, picking up signal from the i.f. output of the front-end tuner, and TR2

is the final transistor feeding the ratio detector transformer. Both transistors are in the common-emitter mode, whereby the emitter is common to both input and output signals, this electrode being "earthed" signalwise by C1 and C2. The transistors are BF115s—npn devices with low feedback capacitances. The low

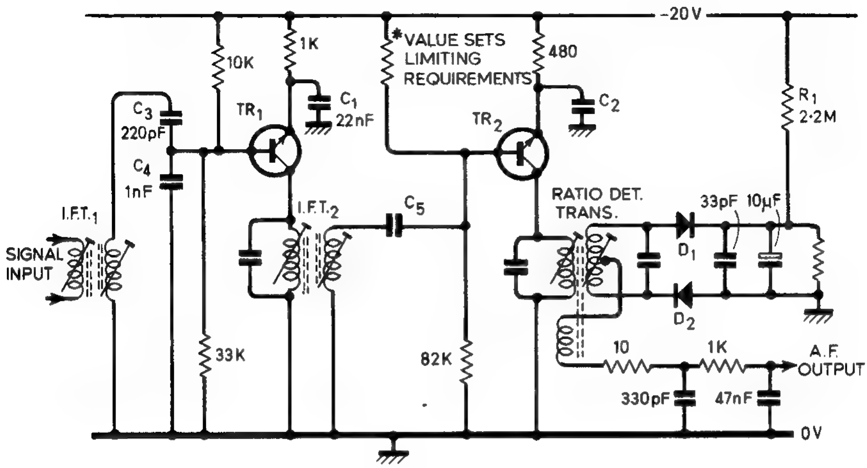


FIG. 5.9. *Two-stage transistor i.f. amplifier feeding ratio detector. The i.f. transformers are designed for critical coupling and the correct impedance is obtained by capacitive-coupling from the secondaries to the bases of the transistors. The detector diodes are given a small forward bias by the 2·2M resistor to the —20V line to improve low-signal linearity.*

feedback capacitances make it possible to secure quite high gain without neutralization, the two stages often being adequate for medium signal strength areas, and with a reasonable front-end the circuit can produce a signal/noise performance of 24 dB when the aerial signal is as low as 5 μV.

The i.f. transformers are designed for critical coupling and are double-tuned. An interesting aspect is the capacitive coupling from I.F.T.1 secondary to TR1 base, via C3 and C4, and from I.F.T.2 secondary to TR2 base, via C5. This sort of coupling is used because of the very low source impedance required at the i.f. (10·7 MHz) which is sometimes more difficult to obtain by inductive tapping. The second stage is designed to "bottom" at high signal levels and thereby act as an a.m. limiter. With a suitable front-end the limiting could be complete when the input signal reaches about 35 μV.

The circuit also shows the f.m. detector, which is an unbalanced ratio detector comprising a pair of OA79 diodes, D1 and D2. It will be seen that the 2·2 M resistor R1 applies a small forward bias to the diodes from the —20 V line, an action which reflects a relatively constant load on to the tuned secondary of the ratio detector transformer with varying signal level. The forward biasing also aids the a.m. rejection attributes of the circuit.

Highly sensitive hi-fi-type f.m. (and a.m./f.m.) tuners often use up to four i.f. stages, with rather special facilities for amplitude-limiting and muting (see, for example, the Goodmans circuit in Chapter 6).

AMPLITUDE-LIMITER

Fig. 5.10 shows a two-stage amplitude-limiter circuit, the base of the first transistor, TR1 being fed from the secondary of the bandpass i.f. transformer I.F.T.1. TR1 is d.c.-coupled to TR2, the d.c. feedback path from the emitter of

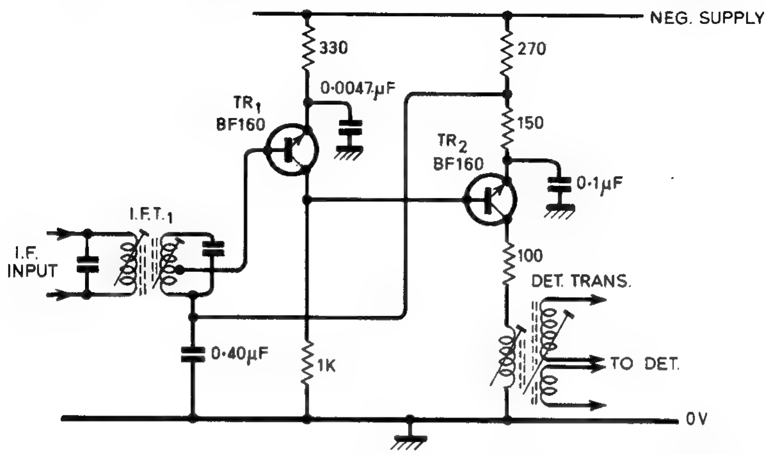


FIG. 5.10. *Two transistor amplifier-limiter system. TR1 is the i.f. amplifier transistor and TR2 the limiter, d.c.-coupling being used in the interests of bias stability and wide-range signal control. In this circuit the correct impedance to TR1 base is obtained by a tapping of i.f.t.1 secondary.*

the limiter TR2 to the base of the amplifier TR1 provides good bias stability and thus maintains a constant collector current in both stages, features which contribute to a constant i.f. signal output over a wide range of input signal levels.

Instead of capacitive coupling to the base of the amplifier transistor TR1, as in Fig. 5.9, the circuit in Fig. 5.10 obtains the correct loading from a tapping on the secondary of the i.f. transformer.

INTER-STATION MUTING

A circuit of an effective solidstate inter-station muting arrangement is given in Fig. 5.11. TR1 and TR2 can be regarded as a switching network operated by a bias applied to the base of TR1. This bias is derived from rectified i.f. signal, the signal being picked up from the emitter of the final i.f. amplifier stage. The signal is rectified by diode D1, and a d.c. potential is developed across R1 making the base of TR1 negative. This cuts off both TR1 and TR2 (since TR2 is d.c.-coupled to TR1) and thus allows the audio from the f.m. detector to go unhindered to the audio output circuit.

However, when there is no signal or a very weak signal at the emitter of the i.f. amplifier, TR1 and TR2 conduct, an action which puts C1 in shunt with the audio path, thereby bypassing the audio to "earth". The correct time-constant is provided by R1, R2 and C2 in TR1 base circuit. The signal level at which the muting circuit operates can be adjusted by the threshold preset in TR2 emitter circuit. When the facility is not required (on very weak signals for instance) it can be switched out by operating the muting on/off switch in TR2 emitter circuit.

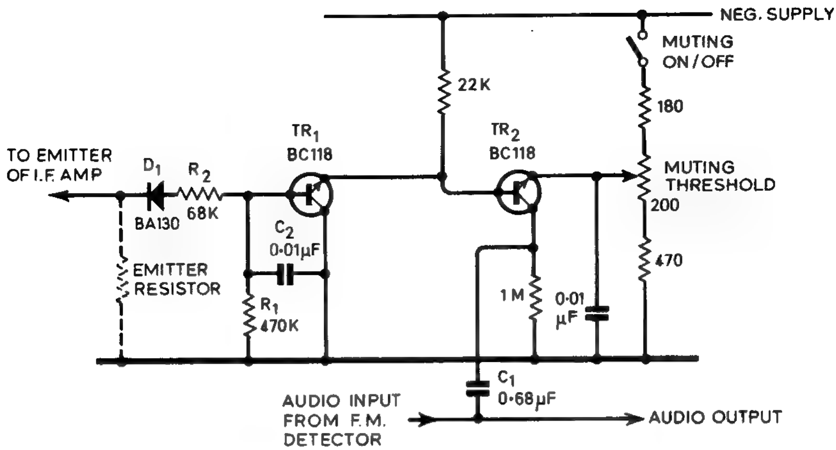


FIG. 5.11. *Inter-station muting circuit in which the two transistors act as a "switch" causing the a.f. path to be shorted out by C1 when the i.f. signal rectified by D1 is insufficiently strong to switch off the transistors.*

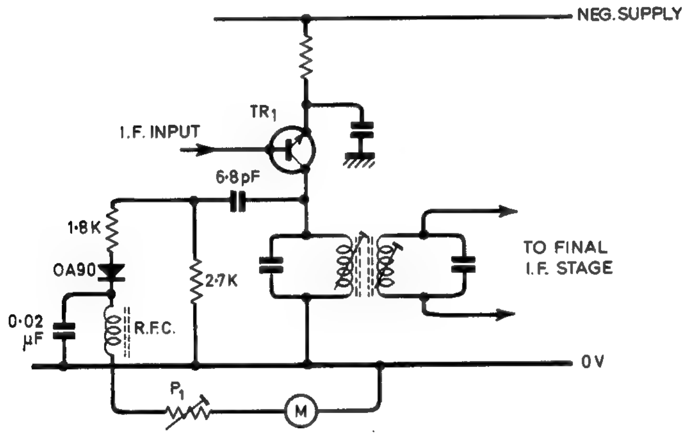


FIG. 5.12. *Tuning indicator meter circuit (see text). Also see Fig. 5.16.*

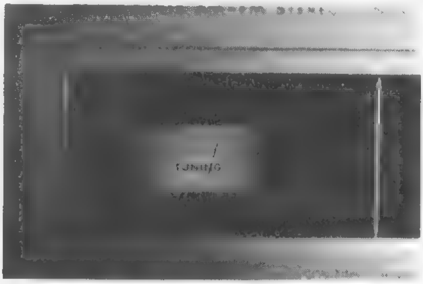
TUNING METER

A typical arrangement used for obtaining an "on tune" indication is shown in Fig. 5.12. Here some of the i.f. signal at the collector of the last but one i.f. amplifier stage is extracted via a 6·8 pF capacitor and 1·8 k resistor and applied to a small semiconductor diode (the OA90 in the circuit). The resulting d.c. is fed into a small moving-coil meter showing on the front panel of the set. When the receiver is accurately tuned the reading will be at a maximum, which is not always the case when the d.c. in the f.m. detector is utilized for this purpose.

The r.f. choke rids the meter circuit of i.f. signal, while the preset P1 allows any full-scale deflection to be established. Such indicators can be very sensitive and provide a good indication of the strength of the aerial signal up to the point of amplitude limiting.

Fig. 5.13 shows a selection of tuning indicators. A very simple arrangement takes the form of a filament lamp or neon bulb which lights to maximum intensity when the station is accurately on tune. This type of circuit is often operated

(*a*)



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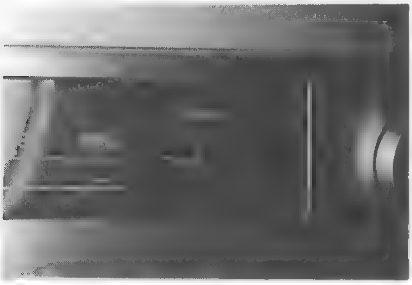


(*c*)

FIG. 5.13. *Types of tuning indicators. (a) bulb-type showing maximum illumination on-tune; (b) moving-coil meter giving maximum deflection on-tune; (c) centre-reading on-tune type.*

through a transistor amplifier stage, giving suitable d.c. gain. F.m.-detector-connected meters are sometimes centre-reading, giving an indication when the detector is accurately "balanced", so to speak, such condition arising when a correctly aligned i.f./detector channel is tuned properly.

INTEGRATED CIRCUIT I.F. CHANNEL

Integrated circuits (i.c.s) used in f.m. receivers are mostly of the so-called monolithic variety, developed on small "chips" of silicon crystal, as distinct from similar devices made on a glass substrate upon which is evaporated thin metallic films, called thin-film circuits. The active and passive elements, like transistors and resistors, are created by etching on the chip, while small-value capacitors are often created from reverse-biased diode elements.

I.c.s make it possible to obtain very high gain and good limiting within a relatively small volume of total circuit. Many i.c.s are encapsulated like transistors and are not all that much larger, yet they contain quite a few transistor elements, diodes and the passive elements just mentioned. They require external circuit elements, like large value capacitors, some resistors and tuned elements, such as i.f. transformers, and the whole circuit is generated on a printed circuit board.

The RCA CA3012 i.c., used in the Truvox FM200IC f.m. tuner, for instance, is a ten-lead package conforming to TO-5 encapsulation. In the Truvox this device is arranged as a wideband i.f. amplifier. It contains no fewer than ten

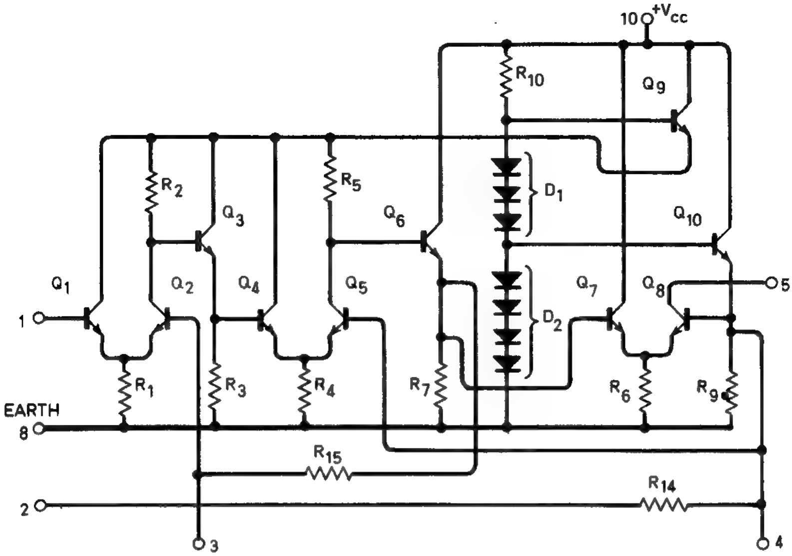


FIG. 5.14. *Circuit of RCA Type CA3012 i.c.*

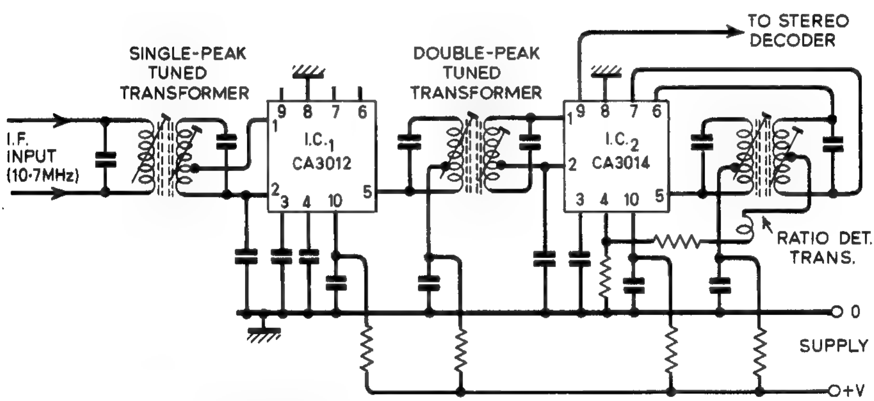
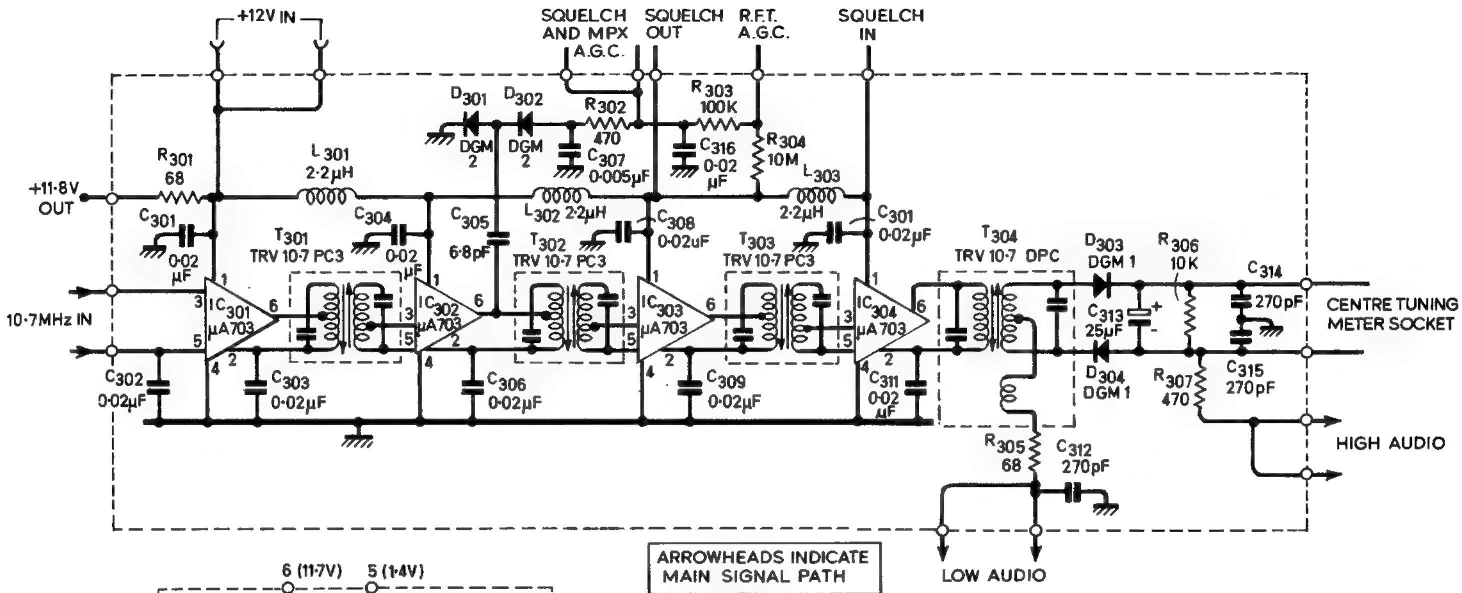


FIG. 5.15. *Basic circuit of the Truvox FM200IC FM tuner i.f. channel using two RCA i.c.s. The ratio detector operates in conjunction with the diodes in the CA3014 on the "average" detection principle owing to the need to use small value capacitors (inside the i.c.) as load reservoirs.*

transistor elements and a multiplicity of diodes (see Fig. 5.14), working out to three d.c.-coupled differential amplifier stages in cascade, an emitter-follower output section and power supply regulation. This device alone has a basic gain around the 65 dB mark at 10·7 MHz, and this needs to be "tamed" and concentrated in the required i.f. bandwidth by the tuned circuits and external components. The Truvox also has a second i.c. (RCA CA3014) which consists of a three-stage d.c.-coupled amplifier/limiter cascode arrangement, power regulating devices, ratio detector diodes and a Darlington pair output stage. The ratio detector of this i.c. is somewhat different from such a circuit based on discrete components, for it will be recalled that normally large-value capacitors are



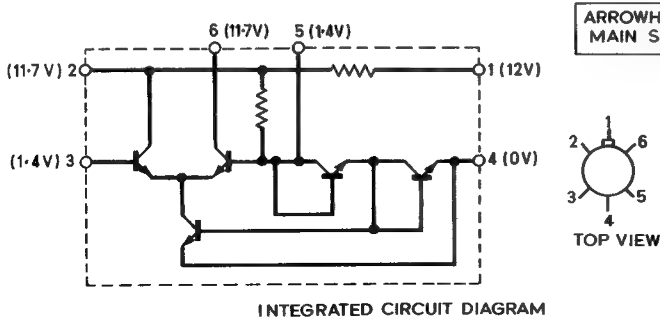


FIG. 5.16. *Circuit of i.c. i.f. channel of the Scott 342-B tuner-amplifier. The circuit of each of the four i.c.s. is given in the inset below. Separate diodes for the ratio detector are used in this circuit, across the balanced load of which is connected a centre-tune meter. The arrowheads are the i.c. symbols.*

adopted to obtain peak rectification from the diode pair; but since capacitors of this value cannot be created in i.c.s, Truvox based the design on "average detection", working in conjunction with a tuned phase-shift transformer of the conventional type, where the load to the diode pair is essentially resistive. The basic i.f. circuit of the Truvox FM200IC is given in Fig. 5.15. I.C.s used in radio circuits of this kind are called *linear* devices to distinguish them from those handling the digital information of computers.

Not all tuners and f.m. receivers use such complicated i.c.s as those in the Truvox. Fig. 5.16 shows the i.c. i.f. channel of the Scott 342-B tuner-amplifier. This uses four less complicated i.c.s as shown by the inset below the circuit. Separate detectors diodes, too, are employed in this circuit.

FIG. 5.17. *This picture shows the small size of the complex RCA CA3014 linear i.c. as used in the Truvox tuner.*

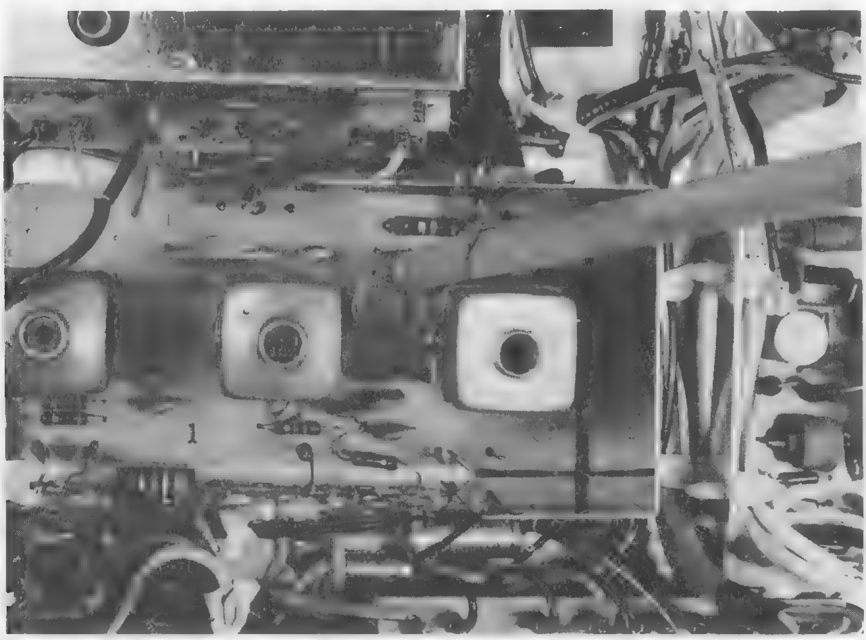


FIG. 5.18. *I.c. i.f. strip. Here the pencil is pointing to one i.c. A second darker one can be seen the other side of the i.f. transformer.*

Incidentally, the centre-reading tuning meter connected across the d.c. output of the balanced ratio detector is here worthy of note.

Fig. 5.17 gives some idea of the overall dimension of the RCA CA3014 i.c., while Fig. 5.18 shows an i.c. i.f. channel in a tuner-amplifier. The pencil is pointing to one such device, while a second can be seen in a black casing a little along the strip.

The inside of a Mullard i.c. is shown in Fig. 5.19. This gives a good impression of the smallness of the silicon chip used in this form of microelectronics.

FIG. 5.19. *Just how small integrated circuits really are is shown by this picture of an unencapsulated Mullard TTL decade counter containing over 120 components. Decade counters are used in computers, desk calculators, adding machines and all types of process control equipment (Mullard photo).*

Some transistor tuners and receivers incorporate a form of a.g.c. with the control being applied to the base of the r.f. transistor to avoid overloading on strong signals. Forward and reverse a.g.c. systems have been used; but with the advent of the f.e.t. input overloading is less of a problem. Instead of using a.g.c., some models have two aerial inputs, one direct and the other via an attenuator (about 20 dB), the latter being used when a local station tends to cause input overloading and crossmodulation troubles. Further information on a.g.c. and input attenuation is given in Chapter 6.

At the time of writing there is a trend towards the use of crystal and ceramic resonators in the f.m. i.f. channel of transistor equipment for selectivity enhancement (see also page 202).

CHAPTER 6

Combined A.M./F.M. Receivers and Tuners

DOMESTIC receivers designed for the reception of both a.m. and f.m. signals often feature a dual i.f. channel which is responsive to a.m. signals at 470 kHz and f.m. signals at 10·7 MHz. A system of ganged switching permits the immediate selection of either a.m. or f.m. by feeding into the i.f. channel either the 470 kHz output from the normal a.m. aerial and frequency-changer section or the 10·7 MHz output from the f.m. tuner unit. The switching system also selects the appropriate detector whose output is conveyed to the receiver's common a.f. stages.

DUAL INTERMEDIATE-FREQUENCY CHANNEL

The network as a whole is cleverly devised and each section well warrants individual consideration. A good start is with the dual i.f. channel.

Probably the chief advantage of the dual i.f. channel lies in its economy of production. From the aspect of a single stage, the arrangement permits the use of the same valve or transistor for both the a.m. and f.m. intermediate frequencies without involving elaborate switching arrangements.

The skeleton circuit in Fig. 6.1 gives the solution to the problem. Here, the i.f. transformers corresponding to the a.m. and f.m. channels are connected in series and wired to the valves in the conventional manner. At first sight it may be thought that such an arrangement could not work efficiently and that one transformer would adversely affect the operation of the other.

This is not so, however, for when the input to the first i.f. amplifier valve V1 is at, say, 10·7 MHz the amplified i.f. signal is fully developed across the f.m. intermediate-frequency transformer because the top ends of the windings, points A and B, are virtually at chassis potential so far as the 10·7 MHz signal is concerned. This is because of the comparatively low total capacitative reactance given by C1 in parallel with the self-capacitance of L1 in series with C5, and C2 in parallel with the self-capacitance of L2 in series with C6. From the i.f. point of view at 10·7 MHz, points A and B can thus be considered as connected to chassis, which is how they are connected in i.f. amplifiers of more conventional arrangement.

When the input to V1 is at 470 kHz nearly all the amplified i.f. signal is developed across the a.m. transformer. This is because the series inductance due to the 10·7 MHz transformer is but a very small part of the inductance due

to the a.m. transformer. Actually, the effect of the inductance of the f.m. transformer tends slightly to reduce the gain of the amplifier, but this is generally a good thing since if the valve is arranged to give optimum gain at the f.m.

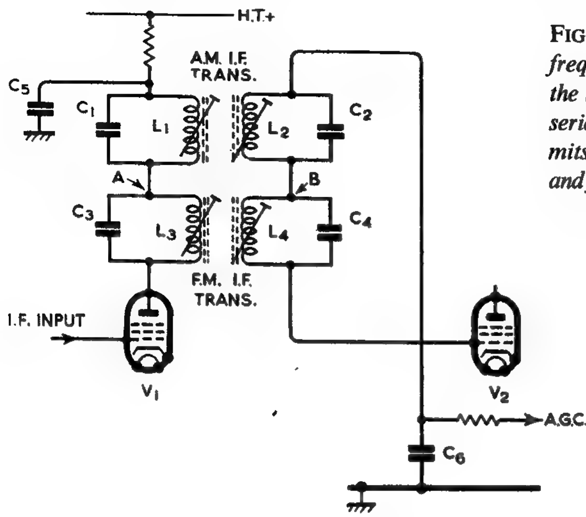


FIG. 6.1. *Because the intermediate frequencies are sufficiently far apart, the transformers can be connected in series without interaction; this permits the use of the same valve for a.m. and f.m. without elaborate switching.*

intermediate frequency a similar gain on the a.m. channel would make the receiver embarrassingly sensitive on this service.

The securing of a balanced output on a.m. and f.m. represents one of the problems in the design of combined receivers, but the effect just described helps to resolve this one without undue difficulty.

While on the subject, there are essentially three other ways of getting rid of excess gain on a.m. Sometimes the a.m. transformers feature a tapping so that only a part of the total signal voltage developed across them is fed to the following stage; resonance of the a.m. intermediate-frequency transformers may be achieved by the use of low L/C ratios; that is, using comparatively high-value fixed tuning capacitors, such as C1 and C2 in Fig. 6.1, and correspondingly low inductance values for the associated coils. This results in considerable reduction of overall gain; it also has the effect of reducing the selectivity of the tuned circuit which is not always a good thing; or sometimes an extra resistor is introduced into the cathode circuit of the i.f. amplifier valve when the receiver is switched to a.m. This, of course, increases the grid bias on the valve and consequently reduces its gain.

Combined receivers using i.c.s may use fewer i.c.s for the a.m. i.f. channel and the full amount for the f.m. i.f. channel. When transistors are used, however, an extra one or two stages may be used for the f.m. i.f. channel, as when valves are used.

In a large number of combined receivers the a.m. and f.m. intermediate-frequency transformers are housed in the same screening can. The a.m. transformer follows conventional design and is invariably tuned by dust cores. The f.m. transformer is designed with consideration to the points already outlined, and is built on its own former by the side of the a.m. transformer.

Fig. 6.2 depicts a transformer of this kind from a combined receiver developed by A. C. Cossor Ltd. The a.m. transformer can readily be distinguished as the one having the greater number of turns, and is the one on the top former in the photograph. The formers are accurately made of a plastic material from a common moulding, and the inside of both sections is threaded to allow screw adjustment of the tuning cores. Also shown on the photograph are the fixed

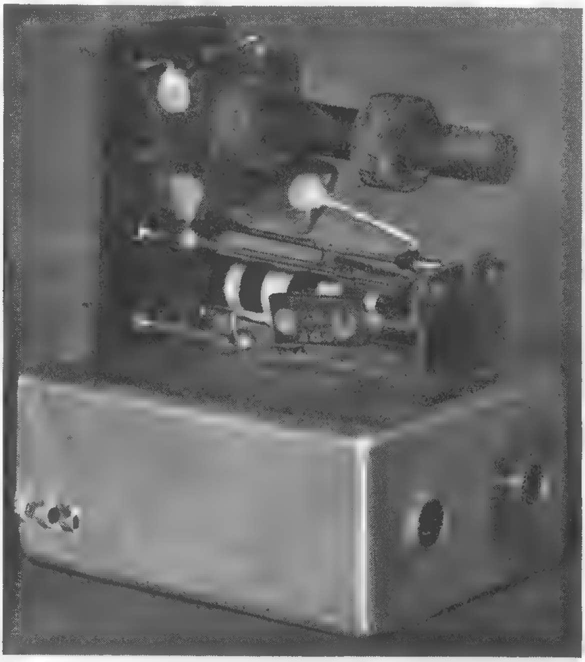


FIG. 6.2. *This illustration of a dual i.f. transformer gives a representative idea of its construction. The a.m. section is shown on the top former.*

tuning capacitors across the various windings. As these are of somewhat critical value, especially those associated with the f.m. transformer, replacement should be made only with the correct value and type of component. The system as a whole does not necessitate screening the two transformer sections, but it is most important for the common screening can to be securely positioned and in good electrical connexion with the receiver chassis, as a means of maintaining a high degree of stability of the i.f. channel.

AN EXTRA INTERMEDIATE-FREQUENCY STAGE

In a large number of combined receivers an extra stage of i.f. amplification is made available on the f.m. channel. How this is accomplished is revealed by the basic circuit of the a.m. frequency-changer section in Fig. 6.3. Here valve V1 represents the a.m. triode-heptode frequency changer, and the tuned circuits associated with the local oscillator section are drawn in block form for simplicity.

As is conventional practice, the two i.f. transformers are loaded at the anode circuit of the heptode section, and a common i.f. output is obtained from across the series-connected secondary windings.

On a.m. the circuit operates normally; the tuned output signal from the a.m. aerial circuits arrives at the signal grid of the heptode by way of the a.m./f.m.

change-over switch (switch 1). Switch 2, being ganged to switch 1, is also in the a.m. position so h.t. is applied to the oscillator section of the frequency-changer valve and removed from the f.m. tuner unit. The 470 kHz a.m. intermediate-frequency signal thus appears across the appropriate i.f. transformer.

On f.m., however, things are a little different; switch 1 picks up the i.f. signal from the output of the f.m. tuner unit and applies it to the signal grid of the

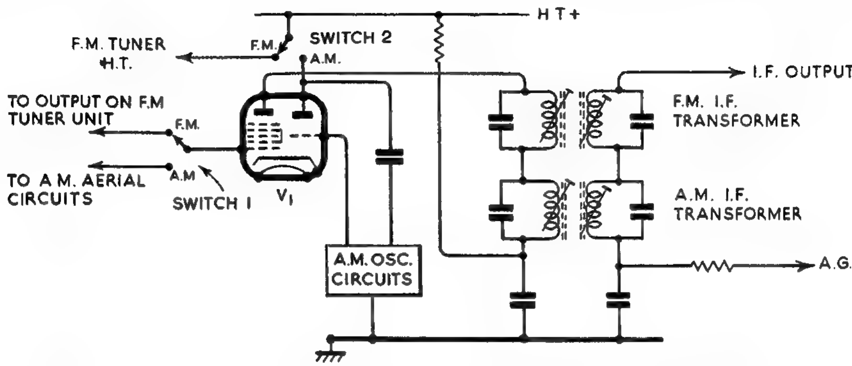


FIG. 6.3. *Basic circuit of the first stage of a combined receiver showing how the heptode section of V1 operates as an extra i.f. stage on f.m.*

heptode, removing the a.m. signal, of course, and switch 2 cuts off h.t. from the a.m. oscillator and applies h.t. to the tuner. Thus, the heptode section of V1 now functions quite well as an extra i.f. amplifier stage, and the 10·7 MHz i.f. signal is developed across the f.m. intermediate-frequency transformer. Either the a.m. or f.m. intermediate-frequency signal is conveyed to the following dual i.f. stage in the usual manner.

As a means of eliminating the possibility of interference, sometimes it is considered desirable to short-circuit one winding on at least one i.f. transformer corresponding to the channel not in use at the time. In the Cossor receivers the unused primary winding of the first dual i.f. transformer is short-circuited by a separate change-over section on the ganged a.m./f.m. switch.

As has already been intimated in previous chapters, combined receivers invariably employ the ratio detector in the f.m. channel. It will now be shown how this circuit makes for ease and economy of production.

The dual i.f. channel is featured right up to the detector stages, thereby enabling each detector to be fed from its own transformer. This method precludes the necessity of switching the detector input circuits, where stray circuit capacitances might well cause trouble, since each detector automatically receives the appropriate i.f. signal by function of the a.m./f.m. change-over system at the front of the receiver, and possibly in the i.f. stages.

Fig. 6.4 shows the circuit of a combined a.m./f.m. detector and a.f. amplifier. Although this circuit is basically representative of nearly all combined receivers, it is actually the one used by A. C. Cossor Ltd. A salient feature of this kind of circuit is the multiple valve V1 (6AK8). It is really four valves in one envelope. There is a single diode with its own independent cathode, two additional diodes and a triode working from a common cathode.

The section of the circuit drawn in heavy line is associated with the ratio detector. The a.m. section of the circuit is drawn normally, while the a.f. and a.g.c. sections are drawn in broken line. This method of presentation is adopted purposely to assist with the identification of the various circuit sections and is not a usual feature of this kind of circuit.

The diode with its independent cathode and one of the other diodes working from the common cathode are used in the ratio-detector circuit. The primary of the f.m. intermediate-frequency transformer is connected in series with the primary of the a.m. intermediate-frequency transformer, so that when the receiver is switched to f.m. the i.f. signal is developed across the ratio-detector transformer.

Resistor R1 and capacitor C1 form the time-constant circuit and are connected in parallel between the cathode of one diode and the anode of the other, as is usual practice with the unbalanced ratio detector circuit—see, for example, Fig. 3.9, Chapter 3.

The a.f. component of the f.m. signal is developed across capacitor C2 and is fed by way of the filter resistor R2 and the a.m./f.m. change-over switch S1 to

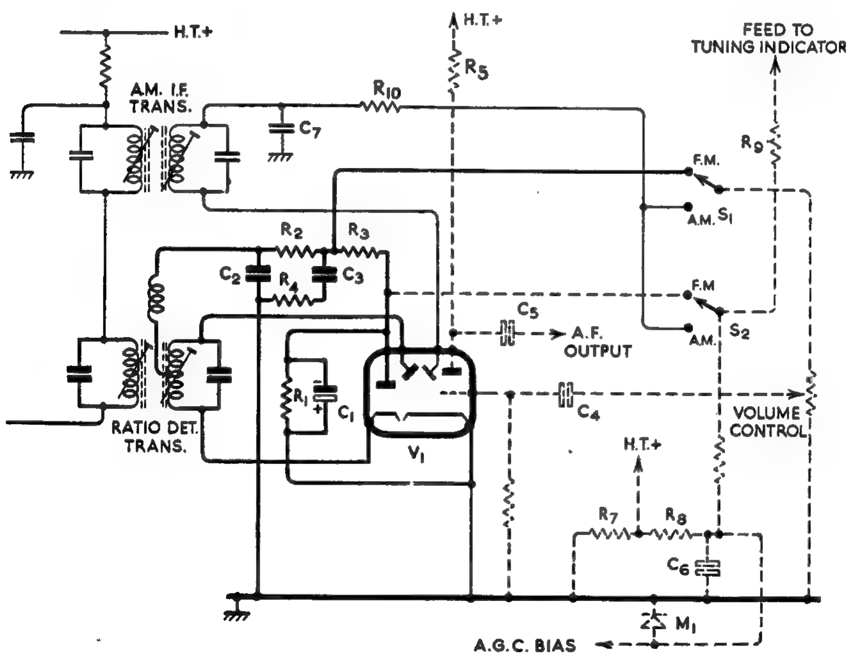


FIG. 6.4. *A combined a.m./f.m. detector and a.f. amplifier built round a triple-diode-triode valve such as the 6AK8.*

the common volume control. The resistor-capacitor series combination R4, C3 gives the correct amount of de-emphasis. The required level of a.f. signal across the volume control is thus passed on to the control grid of the triode section of the multiple valve through the coupling capacitor C4. The signal is amplified and re-developed in the anode circuit across the load resistor R5. From here the

a.f. signal passes through the second coupling capacitor C5 to the final a.f. amplifier or output stage.

ACTION OF THE CIRCUIT ON AMPLITUDE MODULATION

On a.m., switches S1 and S2 change over to the a.m. position and the a.m. intermediate-frequency signal appears across the a.m. intermediate-frequency transformer. The signal is fed to the remaining diode in V1 which serves as a conventional diode detector. Resistor R10 becomes connected in series with the volume control, via S1, and the volume control forms the a.m. detector load. Resistor R10 and capacitor C7 are the a.m. detector filter components.

A.G.C. FACILITIES

Although a.g.c. is a desirable inclusion on a.m. receivers as a means of preventing overloading of the r.f. and i.f. stages due to a powerful local station and of maintaining a reasonably consistent output from a number of stations of varying powers and locations, it really does not apply in the same sense so far as f.m. is concerned.

To prevent overloading of the r.f. and i.f. stages due to a powerful local station it is much better to reduce the aerial signal applied to the receiver by introducing a suitable degree of attenuation than by unduly suppressing the gain of the receiver by the application of a large negative a.g.c. voltage to the controlled valves. Even in receivers featuring a.g.c., aerial signal attenuation is often necessary in locations close to a transmitter. Once the local signal level is known it is a relatively simple matter to apply the required amount of attenuation.

In f.m.-only receivers the amplitude-limiting stage automatically provides for a signal of constant amplitude at the input of the detector. Automatic gain control is sometimes incorporated in such receivers, though it is frequently confined only to the r.f. amplifier stage.

So far as combined receivers are concerned, a.g.c. is also applied on the a.m. channel, but on the f.m. channel it is sometimes switched out of circuit. Since combined receivers often have to operate in fringe areas as well as close to a transmitter, some manufacturers consider that the inclusion of a.g.c. on the f.m. channel is well worth while. Like television signals, v.h.f. frequency-modulated signals are influenced by passing aircraft and weather disturbances, particularly in fringe areas, and rapid fluttering of the signal or slow fading may occur as the result; a.g.c. can assist in reducing these undesirable effects.

Referring again to Fig. 6.4, on f.m. the d.c. potential existing on the negative side of the stabilizing capacitor C1 is used as an a.g.c. bias; the magnitude of this potential depends on the amplitude of the i.f. signal applied to the ratio detector. Since the positive side of C1 is connected to chassis, the polarity of the a.g.c. bias is negative with respect to chassis. Thus, a decrease of signal input, due to fading or other causes, will result in a decrease of negative potential applied to the controlled valve or valves, the gain of the valves will increase and, unless the signal fades beyond the limit of the a.g.c. system, the i.f. signal at the ratio detector will rise to its normal value.

The a.g.c. bias is filtered by resistor R6 and capacitor C6, and the a.g.c. line is returned to chassis through resistors R7 and R8. The positive voltage applied to the junction of these resistors from the h.t. line counteracts the negative potential from the a.g.c. system as the result of a weak signal and thus provides a delay bias so that the controlled valves receive a negative voltage only when the carrier rises above a certain level. This ensures that in fringe areas the receiver is working at full gain, unless, of course, the signal happens to rise above the level of the delay bias.

The metal rectifier M1 serves as a clamp diode to prevent a positive potential from being fed to the controlled valves when the signal falls below the level of the delay bias.

When the circuit is switched to a.m., the negative potential appearing at the junction of R10 and the volume control is used as an a.g.c. bias. The magnitude of this is again dependent on the amplitude of the a.m. signal at the diode detector. It is passed on to the controlled valves in the same way as when the receiver is switched to f.m.

TUNING INDICATOR

In common with ordinary a.m. receivers, f.m. receivers, f.m. adaptors and combined a.m./f.m. receivers invariably feature some form of tuning indicator. Indeed, such indication is even more necessary for f.m. reception than it is for a.m. reception owing to the fact that slight mistuning on f.m. is liable to promote distortion of a much more disconcerting nature than that which may occur as the result of slight mistuning on a.m. Moreover, an f.m. receiver which is apparently correctly tuned, and which gives rise virtually to no distortion at average modulation levels, may introduce considerable distortion when the modulation swings the carrier towards maximum deviation. Such distortion may show up during the louder passages of music in a symphony concert, though being completely absent during the softer passages when the modulation level is relatively low.

In domestic receivers the cathode-ray or magic-eye tuning indicators are the most favoured. These are arranged to work on both a.m. and f.m. channels by tapping in on the a.g.c. line. In the diagram, Fig. 6.4, the control grid of the indicator is connected to switch S2 via resistor R9 so that either the a.g.c. bias

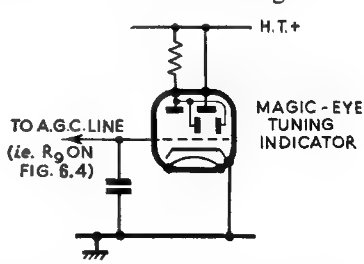


FIG. 6.5. *The circuit of a simple magic-eye tuning indicator which picks up its operating voltage from the a.g.c. line.*

relating to the a.m. signal or the f.m. signal is applied to the indicator, depending on the setting of S2.

Magic-eye tuning indicators represent probably the easiest way of providing an indication of tuning. The circuit is also very straightforward as can be seen in Fig. 6.5. Correct tuning on this circuit is revealed by maximum deflection of the

two images, that is when the "eye" is closed or as near as possible to the closed condition.

There are other methods of connecting the magic-eye circuit to give different effects for the "on tune" condition. One arrangement incorporated in an f.m. tuner design for the home constructor provides an indication of balance of the discriminator. Here the correct tuning is indicated when the eye is just closed; an overlap of the bright areas shows that the receiver is off tune in one direction, while a shadow between the bright areas shows that the receiver is off tune in the other direction. Clearly, an optimum condition of balance is achieved when there is neither a shadow nor an overlap. Circuits of this more specialized nature are usually confined to the more expensive equipment, details of which are given in Chapter 5.

Unfortunately, the simple method of tuning indication is not wholly accurate, since a receiver possessing the right shape of pass-band gives a voltage across the stabilizing capacitor which is essentially constant approximately 100 kHz on either side of the correct tuning point. Thus, the eye remains substantially unaltered over the same frequency range.

In practice, however, a reasonable indication of the correct tuning point occurs as a result of a small peak which is arranged to occur at the centre frequency of the receiver's response curve. It is most important to ensure, therefore, that no spurious peaks are unwittingly created on the response curve on either side of the nominal i.f. during the process of alignment.

A.M/F.M. CHANGE-OVER SWITCH

At very-high-frequency the construction of the circuit and the layout of the valves and components have to be arranged with considerable care so as to maintain good stability coupled with high gain. Since the a.m./f.m. change-over switch is associated with circuits in practically each stage of the receiver, the well known Yaxley rotary switch, which is of almost universal incorporation in a.m. receivers, cannot always be used for wave-band selection owing to the possibility of introducing unwanted couplings between the different circuit sections.

As a means of easing this problem a somewhat unconventional slide-switch has been developed. The switch is constructed of two long narrow sections of low-loss insulating material. One section is secured to the frame of the switch assembly which is fixed to the chassis of the receiver, and the other section is in the form of a slider which can move a limited distance along the fixed section.

Switch contacts are situated at certain points along the fixed section or strip. Each switch usually provides a change-over function, and a group of three contacts are associated with each section. The slider carries small metal strips which engage with the contacts and are so arranged that movement of the slider promotes connexion of the centre contact of a group with either one or other of the outside contacts.

The construction of the switch is such that pressure applied to the slider pushes it in one direction, while a spring pushes it back on release in the other direction. The length of the switch is arranged to suit the particular receiver in which it is

to be employed, and the switch is often mounted so that it extends the entire length of the inside of the chassis from the a.m. frequency changer to the a.f. stages. This enables switch sections to be positioned very close to the circuits requiring switching, thereby obviating the necessity for long connecting leads.

A mechanical coupling device between the spring-loaded slider of the switch and the a.m. rotary wave-change switch permits a single control knob to cater for a.m. wave-change and a.m./f.m. change-over.

In order to give a better understanding of the switching arrangements of a typical combined receiver, a block diagram with the a.m./f.m. switching clearly shown is given in Fig. 6.6. Here S1 connects either the v.h.f. tuner or the a.m.

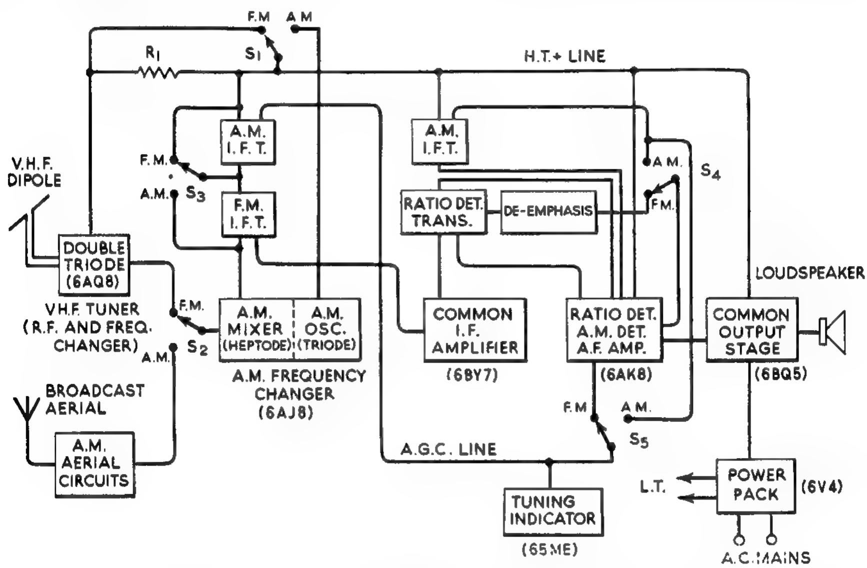


FIG. 6.6. *Block diagram of a typical combined receiver, illustrating the a.m./f.m. switching arrangement.*

oscillator to the h.t. line; S2 selects either the f.m. signal from the tuner or the a.m. signal from the associated aerial circuits; S3 shorts out the unused i.f. transformer primary; S4 selects the appropriate detector; and S5 connects the a.g.c. line either to the ratio-detector stabilizing capacitor or to the a.m. diode load. In receivers not using a.g.c. on f.m., the f.m. switch contact is returned to receiver chassis. In this case, of course, the magic-eye tuning indicator would have to be switched to the stabilizing capacitor.

The purpose of resistor R1 may not be immediately apparent. Actually, its use protects the double-triode v.h.f. tuner valve from "cathode poisoning", and is included to comply with the valve manufacturer's requirements which oppose the operation of high-mutual-conductance valves with their heaters switched on but with no h.t. If such valves are left for a period of time with their heaters alight but with no h.t., an interface is built up on the cathode which eventually destroys the emission. Provided the valve is passing a small current, however, the rate of destruction is reduced considerably. Resistor R1 serves this purpose and also switches off the v.h.f. tuner by reducing its h.t. to about 20 volts.

The block diagram (Fig. 6.6) is representative of the general arrangement of combined receivers where one i.f. stage is available on a.m. and two on f.m. In at least one British receiver, however, three i.f. stages are used on f.m. and yet still only one on a.m. This is made possible by the fact that each detector is fed from its own transformer. Where three f.m. intermediate-frequency stages are required, therefore, it simply means introducing an additional stage between the ratio detector and the second i.f. amplifier. Since this stage is used only for f.m., it can provide amplitude limiting without affecting the performance on a.m.

On f.m. it is also desirable to secure the best possible a.f. response coupled with a low distortion figure. This might well demand the application of a large degree of negative feedback with a consequent reduction of a.f. sensitivity. There is the possibility of counteracting this loss of sensitivity by employing the triode section of the triode-heptode a.m. frequency changer as an extra a.f. amplifier on f.m. It has been seen that this valve section is inactive on f.m., so it could be used in a form of reflex circuit, in association with the a.m./f.m. switching. It may not be possible to get a lot of extra gain from it, though, but sufficient should be available to balance the loss due to feedback.

SOLID STATE COMBINED RECEIVERS

Solid state merely means equipment using the latest semiconductor devices, such as transistors and i.c.s, and at the time of writing this new edition (1969) the trend is towards the so-called tuner-amplifier (also see Chapter 11). A tuner-amplifier is an integration of a quality stereo audio system and an f.m. or f.m./a.m. tuner with decoder for stereo reception (see Chapter 7).

With this kind of equipment the stereo amplifier can form the heart of a hi-fi system, with the inbuilt tuner handling the "radio" requirements. Some models are for f.m. only, while others are designed to take in a.m. too, mostly the medium waveband; though it is possible to obtain versions carrying l.w. and s.w. bands in addition to the m.w. band and f.m. For "quality" reception, of course, only f.m. can be considered, and the tuners of f.m.-only integrations follow closely the lines indicated in Chapter 4, where low-noise bipolar transistors or field-effect transistors deal with the r.f. amplification and mixing and a bipolar device is used for the local oscillator stage.

Many tuner-amplifiers combining a.m. with f.m. use entirely separate front-ends for the two systems, and it is not uncommon to find that the a.m. front-end adopts the straightforward ordinary "radio receiver" techniques, even to the extent of a ferrite rod aerial tuning the first stage as well as collecting m.f. signals, as in transistor portable sets. The f.m. front-end, though, is usually far more sophisticated, following the rules for quality f.m. tuners, some using capacitor-diode tuning and a.f.c.

F.M./A.M. TUNER

However, this section is concerned with combined receivers and tuners, and an interesting circuit of this kind of tuner by Goodmans is given in Fig. 6.7. The stereo decoder part of this tuner is described in Chapter 7 (see Fig. 7.12).

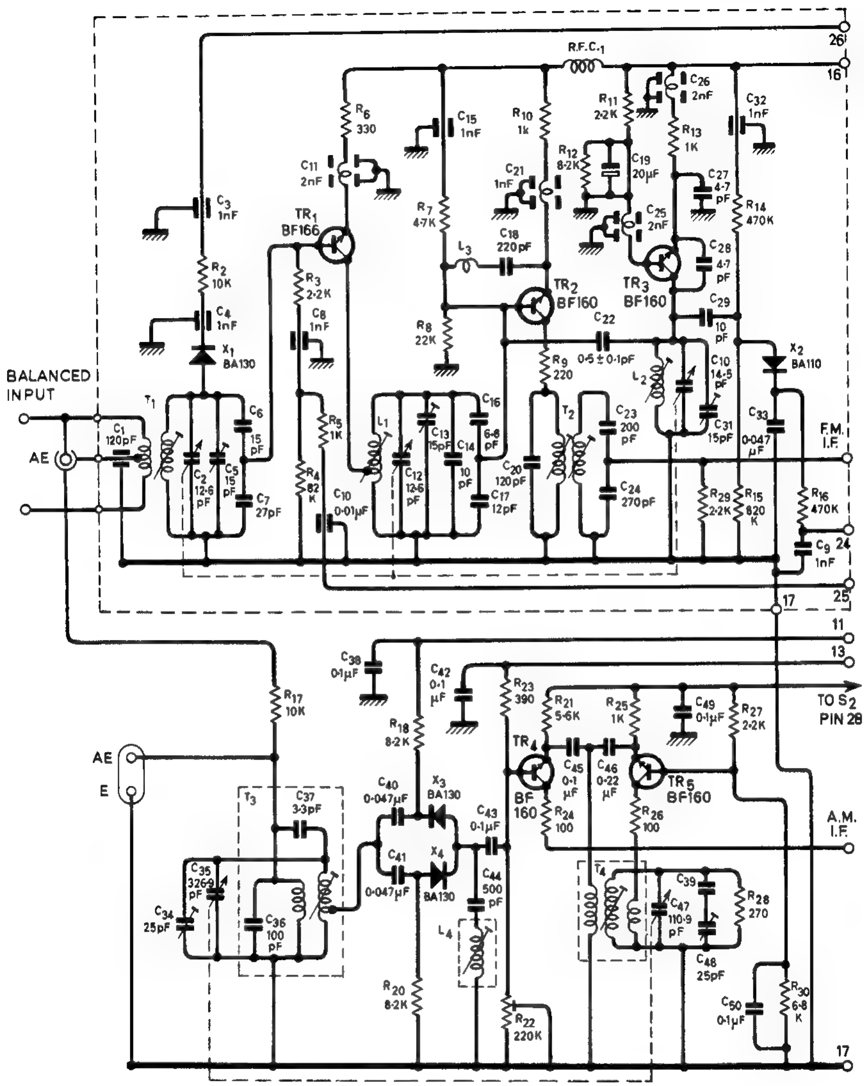
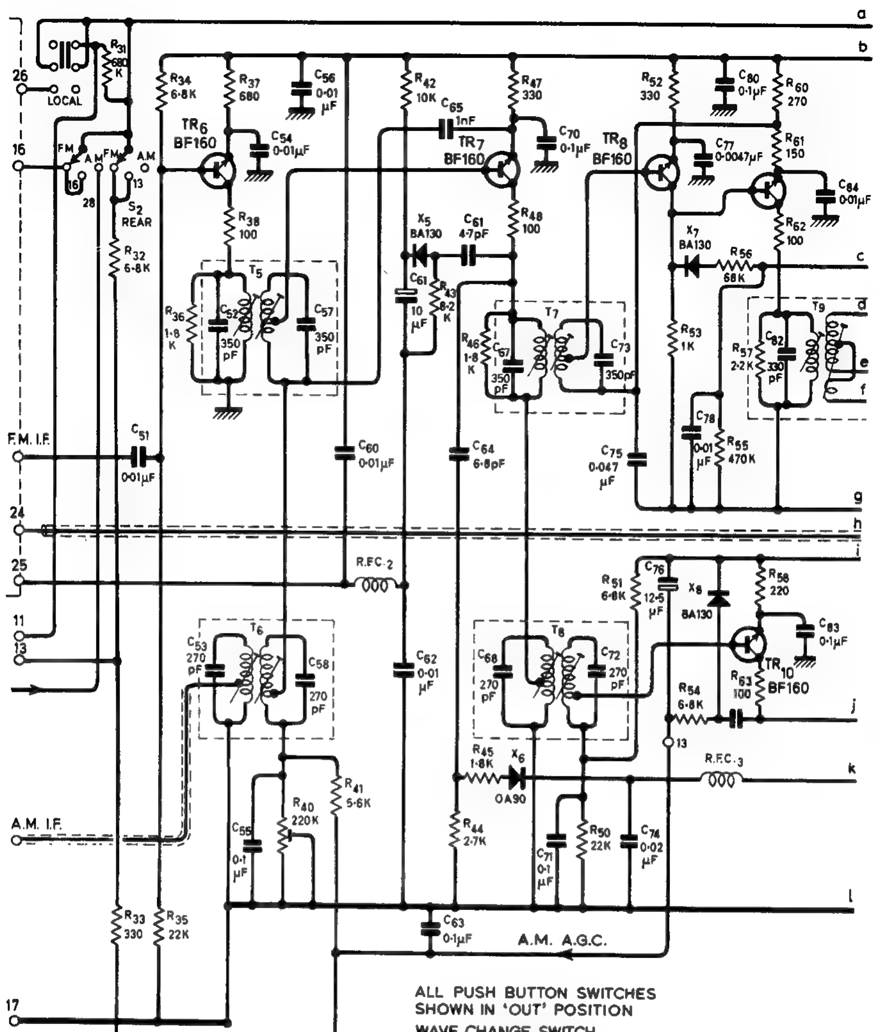
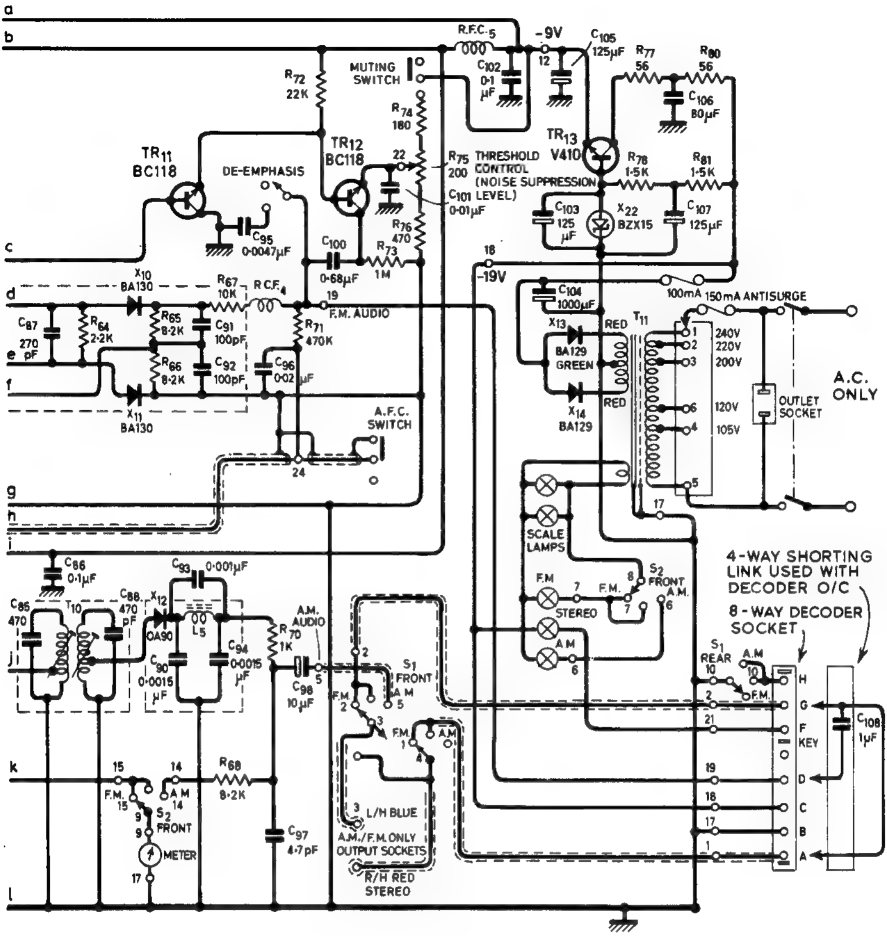


FIG. 6.7. *Circuit of the Goodmans "Stereomax" f.m./a.m. stereo tuner which is exhaustively dealt with in the text. The circuit is continued on pages 108 and 109. The stereo decoder section is considered in Chapter 7 and is illustrated in Fig. 7.12.*



The f.m. front-end employs transistors TR1, TR2 and TR3, acting respectively as r.f. amplifier, mixer and local oscillator, the circuits being tuned by a three-section capacitor gang C2, C12 and C30. Aerial input is designed for 300 ohms balanced feeder but correct working from 75-ohm coaxial is possible either by using a low-loss balun (see *The Practical Aerial Handbook*) or by connecting the feeder screen to the centre-tap of T1 primary and the inner conductor to one of the outer terminals for which a coaxial socket is provided. This action causes the primary to act as a 2:1 ratio transformer, thereby stepping the impedance down by a factor of four—300 to 75 ohms—as well as unbalancing the balanced input.

Capacitive coupling is used from T1 secondary to TR1 base to get the correct impedance and hence the best possible selectivity. TR1 base is biased from the supply line by R3, R4, R42 and diode X5 (the latter two associated with TR7), this stage receiving a.g.c. potential as will be explained later.



Amplified v.h.f. signal is tuned across L1 by C12 section of the gang, and correct matching to the collector is here accomplished by tapping down L1. V.H.F. input to the mixer TR2 base, however, is matched by the capacitive-coupling C16 and C17. The series trap comprising L3 and C18 gives i.f. degeneration, the circuit resonating at 10·7 MHz.

TR3 is put into positive feedback by C28 (see Chapter 4) and the oscillations are tuned by L2 and C30 section of the gang. TR2 mixer base also receives the oscillator signal through the coupler C22, at which point the oscillator signal amplitude is about 90 mV. The local oscillator runs at a frequency 10·7 MHz *below* the signal frequency, and the i.f. signal is developed across the band-pass transformer T2.

A.F.C. is provided by the capacitor-diode X2 across the oscillator tuned-circuit in series with capacitors C29 and C33 which give the necessary "tailoring" to provide the correct ratio of capacitance control. The a.f.c. potential is derived

from the f.m. detector, via the a.f.c. on/off switch. The reverse bias on the diode is about 6 V with respect to chassis.

A.G.C. is provided by a control potential derived from diode X5 picking up signal from the collector of the second i.f. transistor (TR7) and conducting on positive half-cycles. C61 is the "reservoir" capacitor, and the control potential fed back to the base of TR1, which is negative-going with rising signal amplitude, reduces the current in TR1 from 2 mA to almost zero, thereby producing an auto-control of gain approaching 40 dB (see later). This is the *reverse* system of gain control. It is also possible with some of the very latest transistors to reduce the gain by increasing the forward current. That is by making the base go *less* negative in the case of an npn transistor. This is achieved by the use of a resistor of a suitable value in series with the collector tuned load. In this way the collector current rises, the volts drop across the series resistor increases and the collector voltage is reduced, and it is this reduction in collector voltage that reduces the stage gain. This is called *forward* gain control. Forward gain control with the proper type of transistor eases the input overloading problems at high-level inputs.

However, to avoid input overloading, the Goodmans circuit includes a local/distant control. When this control switch is "on" diode X1 in TR1 base circuit conducts and puts a low resistance in series with C4 across T1 secondary, an action which attenuates the input signal by a further 20 dB. The power supply to the f.m. front-end is switched by the f.m./a.m. switch S2.

The a.m. front-end comprises TR4 and TR5. TR4 is the mixer which is also under the influence of the a.g.c. system and which is adjusted by R22 in the base potential-divider network to work at about 2 mA emitter current.

Transformer T3 is concerned with aerial signal tuning in conjunction with C35 section of the gang, and a tap on the secondary winding applies a matched signal to the base of the mixer TR4, via X3 and X4 diode network. C37 gives a little top-capacity coupling to maintain even sensitivity over the m.w. band.

Diodes X3 and X4 are normally conducting and so pass aerial signal without significant attenuation from T3 to TR4 base. However, when the local/distant switch is operated the diodes are reverse biased to some extent and thus become high resistance, thereby attenuating the signal to the mixer transistor. The degree of attenuation is controlled by R31.

The mixer also receives oscillator signal at its emitter from the winding on the oscillator transformer T4 and C45. The oscillator itself (TR5) is arranged in a conventional common-base circuit, with feedback from the collector to emitter, via T4, tuning being by C47 section of the gang. Temperature compensation is provided by C39 in series with the trimmer C48. The a.m. i.f. is developed across the first i.f. transformer T6, the secondary of which is in series with the second f.m. i.f. transformer T5 secondary.

The i.f. channel consists of two stages for a.m. at 470 kHz and three stages for f.m. at 10·7 MHz, TR7 being common to both the a.m. and f.m. i.f. channels.

On f.m. the output of T2 is capacitively tapped—for the correct impedance—to the base of the first i.f. transistor TR6 and the amplified output is developed across the band-pass transformer T5. The signal here is tapped from the secondary to the base of the second stage TR7. TR7 is the second stage and the signal

from this is developed across T7 and then tapped to TR8, which is the third stage. TR9 is the amplitude-limiter whose base is d.c.-coupled to the collector of TR8. This amplifier/limiter circuit is discussed in Chapter 5 (see Fig. 5.10).

Mention has already been made of the a.g.c. diode X5. This conducts on positive half cycles of the signal applied from TR7 collector via C66, the conduction being dependent on the signal amplitude. The diode forms a part of the upper arm of a potential-divider consisting of R42 and R43 relative to TR1 base. Thus, as the conduction increases so the base of TR1 is caused to go more negative. This reduces the current and hence the gain of the r.f. stage in conformity to the reverse a.g.c. system requirements previously described.

TR9 collector feeds a conventional Foster-Seeley discriminator, but with 'phase' coupling winding (Chapter 3), to yield the audio signal. I.F. filtering is by R67 and R.F.C.4, the signal then going to the decorder 8-way socket and thence to the stereo decoder when used (see Fig. 7.12). If a decoder is not used then the 4-way shorting link is plugged into the decoder socket.

TR11 and TR12 constitute the inter-station muting circuit, described in Chapter 5 (see Fig. 5.11.).

When the tuner is used without a decoder (in which the de-emphasis filters are fitted) de-emphasis is provided by switching in C95 which, in conjunction with R67, gives the required time-constant.

Final f.m. points. It is noteworthy that the bottom of T5 secondary is coupled back to TR7 emitter via C65. This ensures that the f.m. i.f. signal appears correctly between TR7 base and emitter. The tuning meter is switched to diode X6 circuit on f.m. so as to indicate maximum signal at TR7 collector when the front-end is tuned correctly. The a.f.c. potential at the f.m. discriminator is switched on and off as required by the a.f.c. switch, which shorts the control voltage to chassis in the "off" position.

On a.m. the 470 kHz signal at TR4 collector is developed across T6 and tapped to TR7 base via T5 secondary. Remember that T5 has virtually no effect on the much lower 470 kHz signal. Bias for TR7 is provided by R41, R54, X8 and R40, the latter being adjusted to give a TR7 current of about 2 mA.

The amplified 470 kHz signal is developed across the band-pass transformer T8 and thence coupled from a secondary tap to TR10 base, which is the second a.m. i.f. transistor, biased by R50 and R51. TR10 collector is tapped into the final a.m. i.f. transformer T10 and tapped out at the secondary to the a.m. detector diode X12. The a.f. filter consists of L5, C90, C93, C94, R70 and C97, the combination of these components producing a fairly steep roll-off in advance of 4 kHz, rising to maximum attenuation at 9 kHz so as to minimize second-channel interference. The d.c. component of the a.m. detector is used to activate the tuning meter in the a.m. position of the meter switch. It will be seen that the f.m./a.m. switch S1 directs the a.m. audio from the detector into both the A and B output channels, which also happens on f.m. when the tuner is running minus a decoder with the shorting link in position.

A.M. a.g.c. is applied to TR4 and TR7 via the base potential divider negative line in conjunction with diode X8 which acts rather like a variable resistor. As the signal increases at TR10 collector, so the signal at X8, fed through C81, causes the diode to conduct more heavily, thereby reducing its forward resis-

tance. This is reflected into the controlled transistor base circuits as an increase in negative bias, reducing the emitter/collector current and thus decreasing the gain. The action is rather similar to that associated with the f.m. a.g.c.

The tuner is powered from a treble-wound mains transformer T11, the primary tapped to suit a wide range of mains voltages, one secondary supplying the various filament lamps (11·5 V a.c.) and the other supplying a bi-phase rectifier (X13 and X14). The rectified output has C104 as the reservoir, and this 19 V d.c. supply is then taken via a 100 mA fuse to the decoder panel (when fitted) direct and to the various other departments via the limiting and smoothing components R77, R80 and C106 and to TR13 base. This is the regulator transistor, the base of which is held at —10 V by the zener diode X22. TR13 and X22 contribute in stabilizing the supply to all the departments, with the exception of the decoder, over a wide range of supply and load variations. The supply to X22 is further smoothed by R78, R81 and C107, while C103 bypasses noise produced by the zener diode. Spurious l.f. and r.f. signals on the supply line are deleted by C102 and C105.

I.C. TUNER-AMPLIFIER

Fig. 6.8 gives the i.c. i.f. channel and a.m. front-end circuit of the f.m./a.m. Lafayette Model LR-500T tuner-amplifier. The f.m. front-end is not shown here but it can be seen in Fig. 4.11 (Chapter 4). Although only three linear i.c.s are shown in Fig. 6.8, there is a fourth in the receiver connected between the output of the f.m. front-end and the first i.c. shown in Fig. 6.8. The first (in the f.m. front-end) is coupled to I.C.1 in series with T206, which is the a.m. i.f. transformer. The f.m. i.f. signal is then developed in the circuit across the band-pass transformers T201 and T202, eventually arriving at the ratio detector transformer T203, progressive signal amplification being given by I.C.1, I.C.2 and I.C.3. The circuit of each linear i.c. is shown in Fig. 6.9, where five npn transistor elements and two resistor elements exist.

A balanced ratio detector is employed, using diodes D204 and D205, with the a.f. obtained from the centre point of the load consisting of R209 and R210. The third winding on the ratio detector transformer T203 in this case is "earthed".

As with the circuits discussed earlier in this chapter, the f.m. and a.m. i.f. transformers are arranged in series, the final a.m. i.f. transformer terminating at the a.m. detector D206. The a.m. filter consists of C210 and R206, with R207 as the diode load. The a.f. is coupled through C212 and is developed across R208, from whence it is fed to the stereo amplifier section via the programme selector switch. The f.m. output is also selected by this switch, but is first passed through the stereo decoder which switches automatically to stereo when the transmission carries a pilot tone (see Chapter 7). The a.f. goes straight through the decoder to both audio channels on mono and separately, after being decoded, to the A and B channels on stereo.

The d.c. component of the a.m. at the junction of R206 and R207 is fed to the base of TR3 through R201. As the signal amplitude varies so does the conduction of TR3, this then being reflected to pin 3 of the controlled i.c.s as a.g.c.

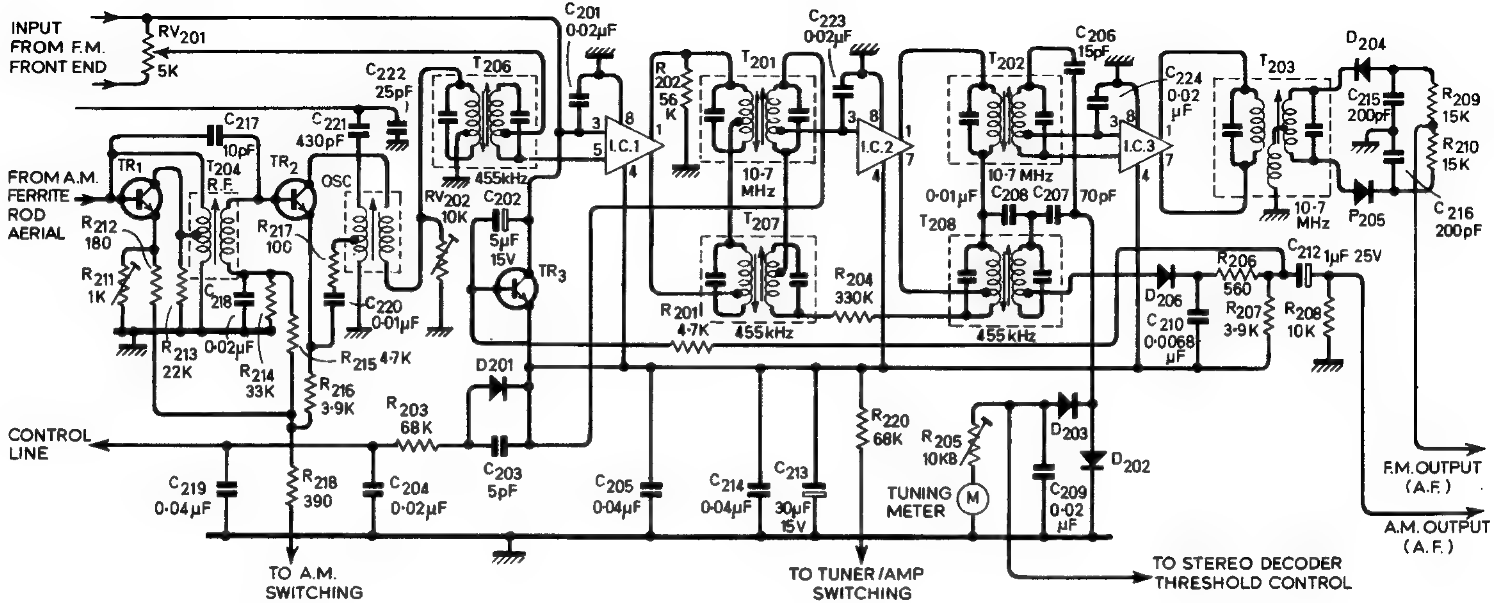


FIG. 6.8. *Circuit of the i.f. channel and a.m. front-end of the Lafayette LR-500T tuner-amplifier. This is fully considered in the text, while the front-end is investigated in Chapter 4, the circuit appearing at Fig. 4.11.*

Looking at Fig. 6.9, it will be seen that the a.g.c. potential is applied to the base of transistor element Q3.

On f.m. D201 provides the a.g.c. potential by rectifying the i.f. signal at the secondary of T201 and conveying the resulting potential to the control line via the filter consisting of R203 and C204.

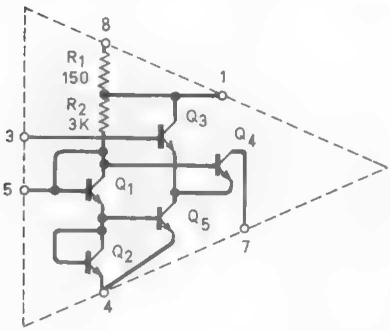


FIG. 6.9. *Circuit of the i.c.s used in the circuit at Fig. 6.8. The combinations of transistor elements can be arranged to give any required signal amplification and control system. It will be seen in the circuit (Fig. 6.8) that I.C.1 receives i.f. signal at the base of Q1 (pin 5) and control potential at the base of Q3 (pin 3). The various schemes used in the other stages of the i.c. channel can easily be worked out from this i.c. circuit.*

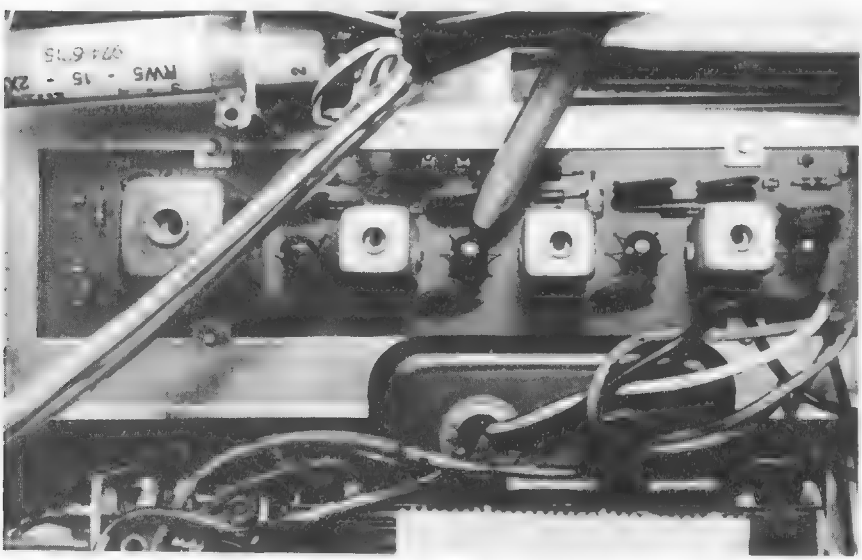


FIG. 6.10. *Four-i.c. i.f. channel. The pencil is pointing to one of the i.c.s.*

The f.m. i.f. signal at T202 secondary or the a.m. i.f. signal at T208 secondary, coupled by C206 or C207, is rectified by diodes D202 and D203 to provide d.c. for operating the tuning meter M. The meter reading can be adjusted by R205 and the potential is filtered by C209. The d.c. here also operates the stereo decoder threshold circuit, with which the circuit sections detailed in Chapter 7 (Figs. 7.13 and 7.14) are associated.

The a.m. front-end consists of TR1 and TR2, the first being the r.f. amplifier receiving signals from a tuned ferrite rod aerial and the second the self-oscillating mixer (e.g., frequency changer). T204 is the r.f. transformer tapped to the

collector of TR1 and core-tuned, while T205 is the local oscillator transformer, also core-tuned. Feedback is between the collector and emitter via the coupling between the two windings in T205. It will be seen that the a.m. i.f. is 455 kHz, as distinct from 470 kHz now commonly adopted in Great Britain. The f.m. i.f., though, is at the "standard" 10·7 MHz.

FIG. 6.11. *Some of the latest tuning-amplifiers employ plug-in transistors of the type here depicted. The three wire ends of the transistor plug into small base sockets of the holder, at which the pencil is pointing. The transistor is shown resting on the tuning drum.*

Another four-i.c. i.f. channel is shown in Fig. 6.10, with the pencil pointing to one of the i.c.s. Fig. 6.11 depicts a section of a tuner-amplifier in which the transistors are of the "plug-in" type. The pencil is pointing to the transistor holder, while the transistor removed from this holder can be seen resting on the tuning drum.

CHAPTER 7

F.M. Radio Stereophony

SINCE writing the first edition of this book a significant development has taken place, whereby it is now possible to reproduce a radio programme on two audio channels simultaneously to obtain the hi-fi stereophonic (stereo for short) effect. This means that a listener with a stereogram or stereo hi-fi system can include stereo radio as a programme source. Indeed, the majority of "compact music systems" and good quality stereograms, as well as hi-fi f.m. tuners, embody facilities for making this sort of reception possible. It is not intended here to dwell on the techniques and aesthetics of stereo reproduction for these aspects are covered in *The Hi-Fi and Tape Recorder Handbook*, by the same publishers; however, we must understand how the system works from the radio point of view; how a mono tuner can be adapted to cater for stereo transmissions; and, of course, the various adjustments required to obtain the stereo effect in its greatest realism.

A stereo system requires two isolated audio channels, one carrying the left-hand information, called the A channel, and the other carrying the right-hand information, called the B channel. The basic arrangement is shown in Fig. 7.1. With ordinary monophonic (mono for short) reproduction a listener is unaware of the directional qualities of the sound; and it also lacks spaciousness. It seems to be coming from a single source, like one loudspeaker; and even if two speakers are used connected to a single audio channel the stereo effect is still not present.

In a concert hall, for instance, the listener is singularly aware of the sound coming from the musicians both individually and as a whole and the spaciousness of the sound; the movement of the sound sources and directivity are other desirable perceptions which cannot be captured by the single channel audio system. However, by the use of two channels arranged as in Fig. 7.1 many of the attributes of the live performance are achieved. In a very well designed system working from a well tailored signal it is not difficult to pick out the various instruments of an orchestra between the two loudspeakers. Care has to be taken to set up the installation properly; but once good stereo reproduction has been experienced there is a great reluctance to return to single-channel mono!

There are two systems of stereo broadcasting. One utilizes two completely independent transmitting channels to carry the information corresponding to

the A and B channels, requiring the listener to employ two receivers; the other employs only one transmitting channel and a single receiver with a special attachment to separate the two lots of information into their respective audio channels. This, of course, is far more economical than the first system since the

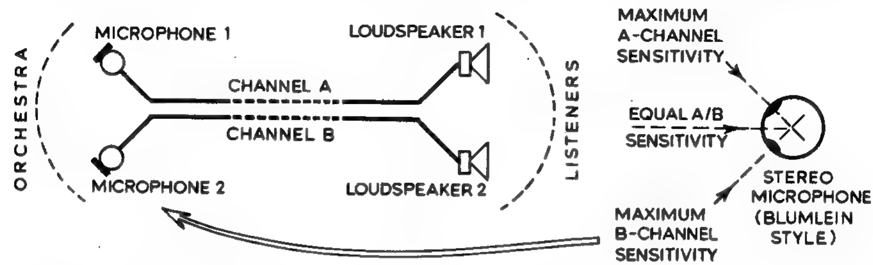


FIG. 7.1. Left: *Basic stereo system with two microphones, two audio channels and two loudspeakers.* Right: *In practice, use is often made of two microphones mounted at the same point one above the other with their axes of maximum pickup arranged at a suitable angle between them, after Blumlein.*

A and B signals are multiplexed at the transmitter and then transmitted as a single, complex signal. Several multiplex systems for stereo radio have been evolved throughout the world, and owing to the shortage of radio space, the cost of transmission equipment and receivers, only the so-called multiplex system could ever be used for domestic stereo by radio.

The system used by the BBC is based on developments by the American General Electric and Zenith Corporations, and it is called the Zenith-GE system. In America, of course, the pre-emphasis given to f.m. modulation is 75 µs, so the system is modified in Great Britain to conform to the 50 µs standard. Moreover, a third audio channel is sometimes used, called a "storecasting channel", in America to provide background music for shops and stores based on a subcarrier of 67 kHz; but this is not utilized on the British system.

As we have seen, a two-channel stereo system initially employs two microphones. These provide the A and B signals, which are audio ones. Clearly, then, if the A and B signals are added together we would obtain an audio signal representative of a mono signal derived by a single microphone placed midway between the A and B microphones of a stereo system. By similar reasoning, the subtraction of the B signal from the A signal (giving A — B) would yield an audio signal which contains all the information relating to the stereo effect. We can thus call the A + B signal the M (M for mono) signal and the A — B signal the S (S for stereo) signal.

The M and S signals are created at the transmitter by a device called a *matrix*, which both adds and subtracts the A and B signals fed to it. We shall see later that a similar piece of network at the receiver—in the stereo decoder—does virtually the opposite to that at the transmitter. It accepts both the M and S signals and by correct addition and subtraction produces the isolated A and B signals which are communicated to separate audio sections for driving the left and right loudspeakers.

The block diagram in Fig. 7.2 shows how the system works at the sending end.

The A and B signals are sent to the matrix through the British 50 µs pre-emphasis filters and the resulting M signal is passed on to the main modulator of the transmitter through a delay network so that this signal arrives at the main modulator at the correct time relative to the other signals fed to it.

The S signal goes to a double-sideband amplitude modulator which also receives a signal at 38 kHz. This is called the *subcarrier* signal, for it is upon this

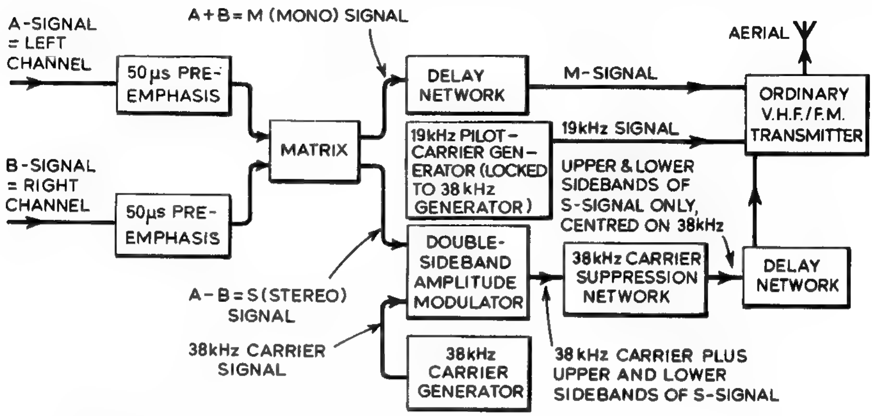


FIG. 7.2. *Basic matrix pilot tone stereo system as it applies at the transmitter.*

that the S signal is amplitude modulated. Such a modulation results in the production of lower and upper S-signal sidebands; but since the 38 kHz subcarrier proper would tend to restrict the depth of modulation of the other, more important, components of the multiplex signal, the subcarrier is suppressed and only the S-signal upper and lower sidebands are applied to the main modulator; again, after passing through a corrective delay network.

Now, since the subcarrier is suppressed, "unscrambling" of the multiplex signal at the receiver's decoder demands that the subcarrier is reformed in accurate frequency and phase, and for this to happen some sort of "synchronizing signal" must be available, the parameters of which are locked to those of the subcarrier. This requirement is provided by a so-called *pilot tone* signal which is fed to the main modulator along with the other signal components mentioned.

The synchronizing subcarrier, however, is arranged to have a frequency exactly half that of the subcarrier, and since the subcarrier is at 38 kHz the *pilot tone* is at 19 kHz. In actual fact, the subcarrier is derived from the pilot tone generator at the transmitter by a frequency-doubler circuit, and in this way the two signals are related.

Of course, if the pilot tone were passed to the modulator at high amplitude there would be no point in going to all this trouble; but in practice this signal has a very small amplitude and thus occupies only a small part of the total modulation depth, ± 75 kHz for 100 per cent modulation as earlier chapters showed. We are not particularly concerned with the amplitude of the pilot tone provided it can over-ride the noise of the system at the receiver, for amplification is usually applied at the decoder. Its amplitude is, in fact, 9 ± 1 per cent, leaving about 90 per cent available for the M and S signals.

The main modulator, therefore, receives a multiplex signal comprising the M, S and pilot tone components, and their spectrum and relative modulation depth are shown in Fig. 7.3. The instantaneous deviation of the main carrier can be expressed as:

$$0{\cdot}9(M/2 + S/2 \sin 2\omega t + 0{\cdot}1 \sin \omega t),$$

where the first term represents the main mono signal, the second the suppressed-carrier subchannel and the third the pilot tone (where $\omega/2\pi$ equals the 19 kHz

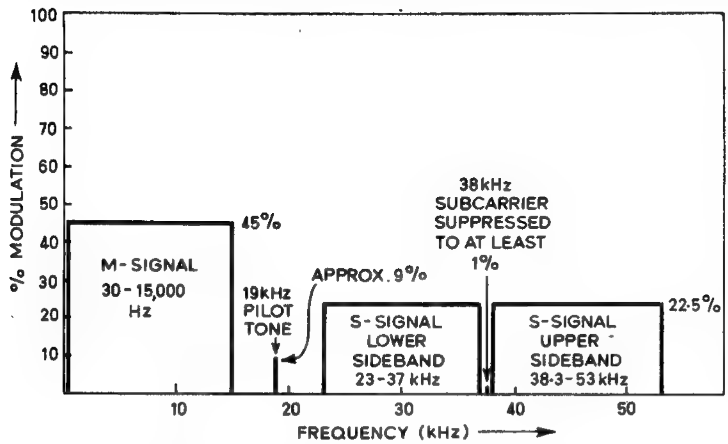


FIG. 7.3. *Frequency spectrum of the multiplex signal showing modulation depth. When the M-signal modulation depth increases the depth of the S-signals decreases, and vice versa. The subcarrier is suppressed to less than 1% modulation depth and this with the pilot tone accounts for about 10% modulation depth. The remaining 90% of the modulation depth is thus divided equally between the maximum A and B information. The total modulation depth in the diagram amounts to 100% made up of 45% of the M-signal, 22½% of* each *S-signal (upper* and *lower sidebands) and 10% of the pilot tone plus residual subcarrier. When the S-signals equal zero, therefore, the M-signal can rise to an amplitude corresponding to 90% modulation depth.*

pilot carrier frequency). The first two terms used 90 per cent maximum of the available modulation depth (where in the British system ± 75 kHz corresponds to 100 per cent modulation) and the third and residual subcarrier take up to 10 per cent. The first two terms reveal that the S information is included in the signal without a limit being imposed on the M information.

Any ordinary f.m. set (mono) thus receiving a stereo-encoded transmission will yield merely the mono information—the M/2 part of the above expression. The remainder of the f.m.-detector signal, being well above audio frequency, is deleted by the de-emphasis network. In a stereo receiver, therefore, or a mono set with the addition of a decoder, the de-emphasis needs to be removed from the detector, thereby allowing the signal components above 15 kHz to provide the "second channel".

It follows, therefore, that a set making use of the M/2 part of the signal will only have available 90 per cent of the mono deviation, which means that the signal/noise performance could suffer by about 4 dB if the aerial signal strength is only just about sufficient to bring the set to maximum noise performance

threshold on an ordinary straightforward mono transmission. When the whole signal complex is exploited in a stereo system, the noise performance (relative to mono under similar minimum aerial signal conditions) could depreciate by as much as 20 dB. It is imperative, therefore, for a stereo system to be in receipt of the maximum aerial signal as is possible to obtain in the area, even though this may mean the employment of an aerial of significantly higher gain (and height) than used previously on a mono system.

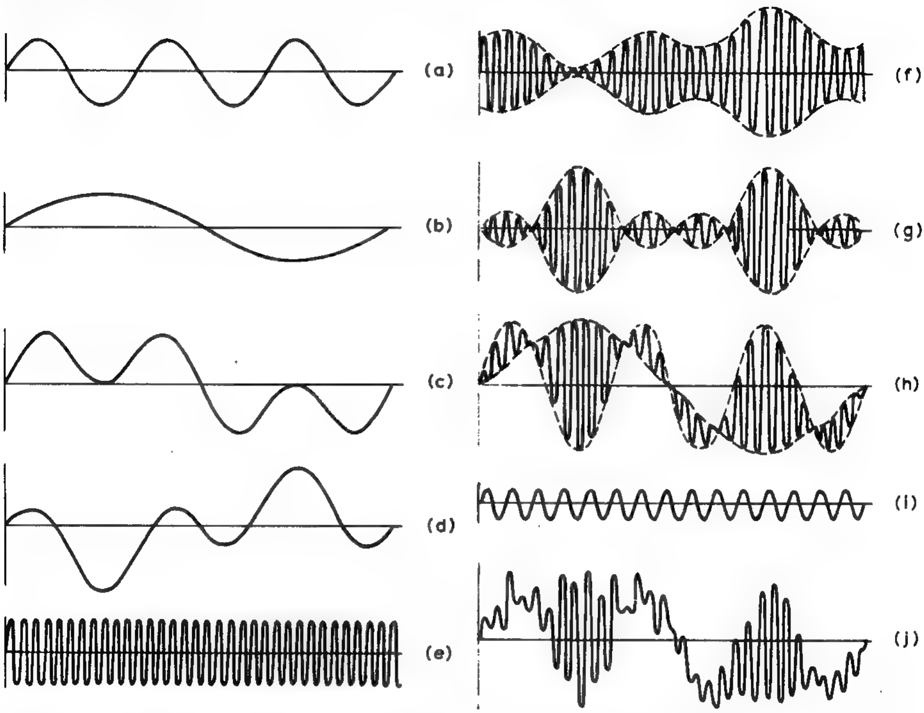


FIG. 7.4. *Signal components (a) to (i) eventually form the multiplex signal at (j). All these signals are described in the text.*

The pilot tone system of stereo broadcasting is thus compatible, meaning that a stereo-encoded signal will work a mono set in perfect mono balance without alterations of any kind. It is the same in this respect as colour television, where a colour-encoded signal will give a good monochrome picture on a black-and-white-only set. In colour TV we also have so-called *reverse compatibility*, where a colour set will yield monochrome pictures from a black-and-white transmission. There is also reverse compatibility in stereo broadcasting, for many stereo decoders and stereo-ready sets will give correct mono reception from a mono-only transmission, the decoder switching automatically between stereo and mono, according to whether the station tuned is stereo-encoded or plain mono. Indeed, many sets and tuners carry a "stereo beacon" (indicator) which illuminates when a stereo-encoded transmission is tuned in, this being activated by the pilot tone signal.

The way that the signal complex is created in terms of waveforms is illustrated in Fig. 7.4. The A and B audio signals are shown at (a) and (b) respectively.

(c) shows the M signal and (d) the S signal (the latter obtained by the graphic superimposition of (a) and (b)). The 38 kHz subcarrier signal is shown at (e), while (f) shows the subcarrier amplitude-modulated with the S signal at (d). Waveform (g) results from the suppression of the subcarrier from the signal at (f). When the M signal is added to waveform (g) we get waveform (h). (i) is the pilot tone signal and when this is added to (h) we get the final complex or multiplex signal shown at (j).

To summarize, therefore, we have seen that the f.m. detector carries a spectrum of signals as shown in Fig. 7.3 when the de-emphasis network is removed, and when it is fed with a stereo-encoded input, and that the signal waveform is after the style of that in Fig. 7.4 (j) when the A and B signals are sine waves of the approximate frequency difference as revealed at (a) and (b) in Fig. 7.4.

It now remains to see how the multiplex signal is processed eventually to deliver the original A and B audio signals. There are three schemes available for decoding based on (i) the *matrix decoder* (rather the reverse of the matrix encoding earlier considered), (ii) the *switch decoder* and (iii) the *envelope decoder*. These can differ quite a bit in detail in practice, though the net result is the same.

BASIC MATRIX DECODER

Fig. 7.5 shows the matrix decoder in block form. The M signal is recovered from the set's f.m. detector by filter No. 1, passing signals from 30 to 15,000 Hz, while the S-signal sidebands are passed to the a.m. detector through band-pass filter No. 2, operating from about 23 to 53 kHz. Filter No. 3 is tuned to the 19 kHz pilot carrier frequency, which is amplified and doubled to reform the suppressed subcarrier signal, this then being applied to the synchronous detector

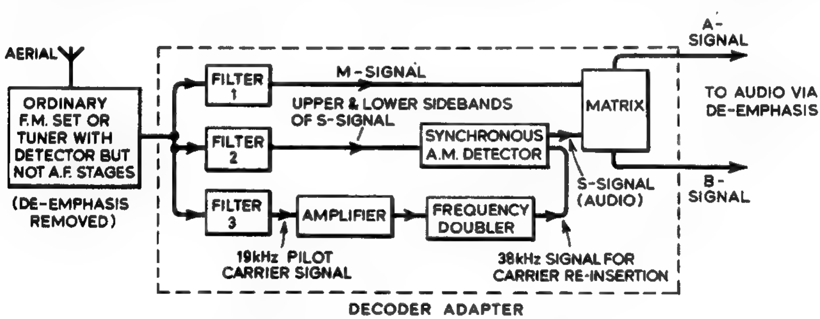


FIG. 7.5. *Block diagram of "matrix" decoder.*

with the S-signal sidebands. The a.m. of the S-signal is thus demodulated, and the audio components of that signal, along with the M-signal, also audio of course, are fed to a matrix containing a "subtractor" and "adder", the action of which yields the original A and B signals. What happens is that the "adder" processes (A + B) + (A — B) to give 2A and the "subtractor" processes (A + B) — (A — B) to give 2B. De-emphasis is applied to each channel before application to the audio circuits.

It is noteworthy that synchronous a.m. detection can only occur properly when the reformed subcarrier is phase-coincident with the subcarrier suppressed

at the transmitter, so due attention has to be paid to the phasing of the doubled 19 kHz pilot tone relative to the S-signal sidebands. Sometimes the de-emphasis is applied to the M-signal prior to its introduction to the matrix. For crosstalk compensation a network composed of a pair of ganged preset resistors connected between the S-signal input to the matrix and the A and B signal outputs is not uncommonly incorporated. This regulates the amplitude of the difference signal relative to the separate A and B outputs, so when properly "balanced" tends to cancel out the A signal in the B channel and the B signal in the A channel.

BASIC SWITCH DECODER

The switch decoder in block form is shown in Fig. 7.6. One way to look upon the stereo system is as a scheme which alternately switches the A and B channels simultaneously at the transmitter and receiver. The switching rate is based on the 38 kHz subcarrier frequency, with synchronism effected by the pilot tone.

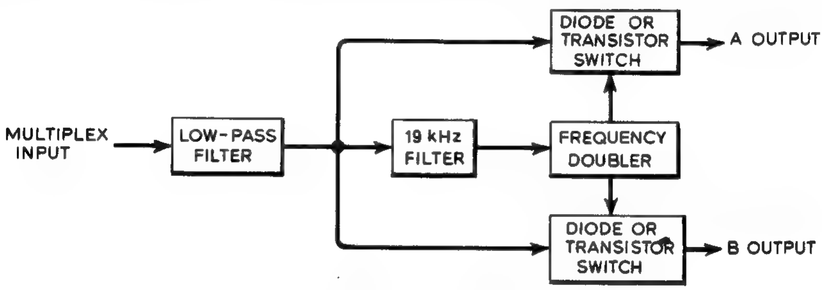


FIG. 7.6. *Block diagram of switch decoder.*

Thus at one instant in time the A channel at the transmitter is switched on, as also is the A channel at the receiver, thereby allowing the A channel only to be reproduced in the left-hand loudspeaker. One thirty-eight-thousandths of a second later the A channel closes down and the B channel opens, again at the transmitter and set in synchronism. This brings in the right-hand loudspeaker. The switching rate is so rapid that one gets the impression that the signals in the two channels are perfectly constant and isolated. The switching frequency is well above hearing, of course. This way of visualizing the system reveals the great importance of accurate synchronism, and gives some idea of what would happen if the receiver switching failed to coincide with the transmitter switching.

Returning now to Fig. 7.6, the low-pass filter through which the multiplex signal from the f.m. detector is passed is limited to the top end of the spectrum (about 53 kHz—see Fig. 7.3) so that all the signal components are fed to the A and B electronic switches. However, the pilot tone is separately filtered through a 19 kHz circuit from whence the signal is taken to a frequency-doubler, giving the 38 kHz subcarrier locked to the pilot tone which, already being locked to the subcarrier at the transmitter, yields the correct synchronism.

The synchronized subcarrier is then used to drive the two counter-phased electronic switches, and since these are also in receipt of the complete stereo signal, the A and B signals are channelled to the appropriate outputs as already explained.

ENVELOPE DECODER

One way of achieving envelope detection is shown by the block diagram in Fig. 7.7. The multiplex signal from the f.m. detector is fed to an amplifier which is designed to yield both the pilot tone signal and the complete stereo signal at separate outlets and at correct impedances. A matrix ("summing" circuit) is

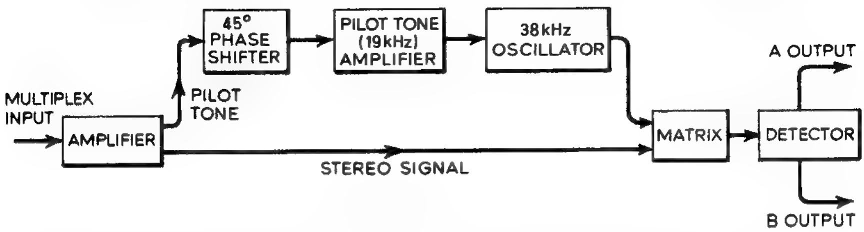


FIG. 7.7. *Block diagram of envelope decoder. This and that in Fig. 7.6 are based on time division multiplex.*

used at the end of the system to feed the envelope detector proper, the matrix itself being fed with correctly phased subcarrier signal and stereo signal through the two circuits shown. The correct phase-shifting is achieved by the first block in the pilot tone channel, the pilot tone then being amplified. However, instead of being doubled to produce the subcarrier, in this case the subcarrier is generated by a 38 kHz oscillator, the signal from which is locked in phase by the amplified 19 kHz pilot tone. It is possible, of course, to double the pilot tone frequency as before to obtain the subcarrier; but the scheme shown in Fig. 7.7 is not uncommonly employed in practical circuits. Another variation of the theme lies in the use of a *19 kHz oscillator*, the output of which is synchronized by the pilot tone signal and *then* doubled to give the 38 kHz subcarrier signal.

The action of the matrix and detecting end of the system is illustrated in Fig. 7.8. The stereo signal at (c), derived from the A and B signals at (a) and

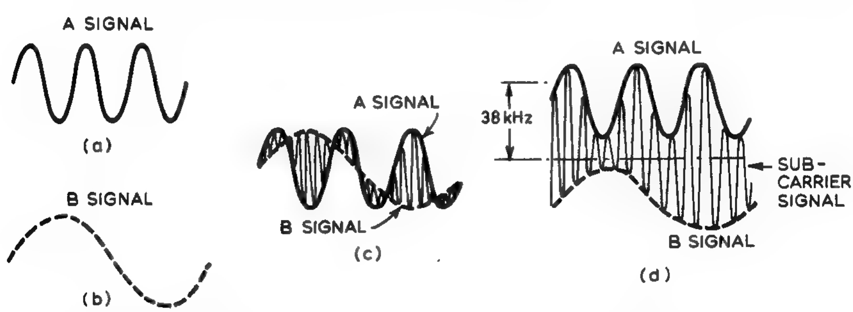


FIG. 7.8. *These waveforms reveal the basic action of time-division multiplex. (a) and (b) show the A and B audio signals respectively, (c) shows the two signals in multiplex, while (d) shows how the two "envelopes" (of the A and B audio signals) are effectively separated by the subcarrier. See also the waveforms in Fig. 7.4.*

(b), is composed of two modulation envelopes overlapping each other. Now, when the 38 kHz subcarrier signal is added the two envelopes are pushed apart, so to speak, as shown at (d), so that one envelope corresponds to the A signal (a)

and the other to the B signal (b). This action occurs in the matrix, and it is then the job of the detectors to demodulate the envelopes separately to produce the original A and B signals.

There are various ways of achieving this sort of matrix-detection, using two diodes, four diodes or even two transistors, the latter in particular operating rather like the electronic switches of Fig. 7.6. In essence, however, one half of the detector system switches in synchronism to demodulate, say, the A signal envelope, while the other half switches to demodulate the B signal envelope.

To sum up, therefore, decoding is based on the synchronous detector principle by matrixing, switching or envelope detection, and practical circuits sometimes integrate parts of the three systems to produce the isolated A and B outputs as efficiently as possible and with the least cross-talk. As we have seen, the subcarrier can be reconstituted by doubling the pilot tone frequency, by doubling a 19 kHz oscillator signal locked by the pilot tone or by locking a 38 kHz oscillator signal by amplified pilot tone signal. The most popular scheme would seem to be doubling the frequency of pilot tone signal to yield the 38 kHz subcarrier; but there are a few decoders that actually employ 38 kHz or 19 kHz oscillators.

Envelope detectors and switching decoders, when the former use two diode pairs and the latter transistors (or two diode pairs), are automatically compatible for a non-stereo transmission; however, when there are only two diodes in the circuit, which is rare these days, the decoder needs to be switched out of circuit when the transmission is without a stereo pilot tone. It is the pilot tone or its doubled 38 kHz signal, therefore, that switches the diodes to give audio-frequency continuity to the A and B channels on a stereo transmission; no signal continuity being provided without the pilot tone switching signal.

Nevertheless, some stereo tuners and decoders embody a stereo/mono switch which, in the "stereo" position, provides for automatic switching between stereo *and* mono—whichever happens to be tuned in—with the former type being indicated by a so-called "stereo beacon" (referred to in greater detail later), worked from the pilot tone, illuminating.

In the "mono" position the decoding action is effectively muted, the decoder then delivering signals in the mono mode only, on both mono *and* stereo transmissions. The idea of this switch is to allow the best possible signal/noise performance from a weak tranmission that is stereo-encoded, and that is too weak to operate the decoder adequately. Under such a condition the background noise can rise to a very high level indeed when the system is switched to "stereo", dropping by 20 dB or so, depending on the aerial signal strength, when the switch is changed to the "mono" position.

When the decoder employs a 19 kHz or 38 kHz oscillator, then this switch mutes the oscillator in the "mono" position, again significantly improving the signal/noise performance when the stereo-encoded signal is weak. It has already been mentioned that an aerial signal some 20 dB above the minimum required for mono is necessary to secure a comparable signal/noise performance from a stereo transmission.

Some decoders also employ a "phasing control" which, when a 38 kHz bandpass filter is switched into the stereo signal feed to the matrix or detectors, is adjusted by listening for maximum sound output in the subcarrier signal. The

band-pass filter lets through only the subcarrier, heavily attenuating the M and S signals, so any residual audio here will then easily be heard. The phasing is correct when the control is set for the loudest audio signal.

CROSSTALK REDUCTION

Crosstalk between the A and B audio channels is at a minimum when the least A signal breaks into the B channel and the least B signal breaks into the A channel. Poor crosstalk performance can result from incorrect phasing of the reconstituted subcarrier relative to the S signal sidebands, from poor bandwidth of the receiver's i.f. channel and from phase distortion therein. Some crosstalk also arises from the nature of the decoding itself, even when the other factors just mentioned are satisfactory.

Some decoders, therefore, incorporate so-called crosstalk compensating circuits, one such arrangement having already been mentioned under the heading "Basic Matrix Decoder" on page 120.

Crosstalk in the switch and envelope detectors tends to result from the nature of the 38 kHz switching signal. A rectangular wave with a 1 : 1 mark/space ratio is normally used for switching time division multiplex systems, of which the switch and envelope detectors or decoders are examples; but when a sine wave switching signal is used the system fails completely to suppress the A channel signal when the B channel is being sampled and the B channel signal when the A channel is being sampled. The fundamental reason for this is illustrated in Fig. 7.9, the main waveform here representing the modulation of one channel

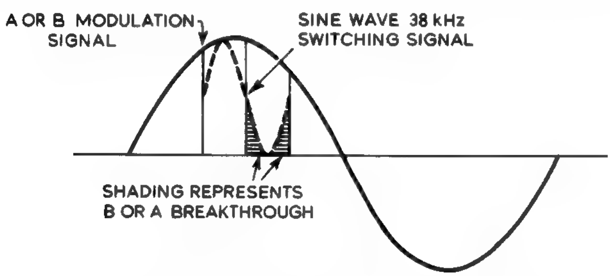


FIG. 7.9. *The basic cause of crosstalk is shown here on one channel only. The other channel is considered as being at zero and, for clarity, no pilot tone is shown. Crosstalk compensating circuits, using one or two preset controls, serve to reduce the crosstalk by introducing neutralizing signals of opposing phase (see text).*

only—the other channel being at zero and the pilot tone signal removed. It will be seen that when the quasi-sine wave 38 kHz switching signal (e.g., the sub-carrier derived at the decoder) samples the modulation it leaves a residue of signal during the period that the other channel's modulation is being sampled, and this is the crosstalk signal. Perfectly rectangular 38 kHz switching pulses would improve matters, but such pulses cannot be obtained via a system whose bandwith extends only to about 53 kHz. Indeed, it would require a bandwidth of 300 kHz, at least, to handle 38 kHz rectangular pulses without rounding.

Crosstalk compensation, based on the addition of attenuated and inverted M signal to the resulting crosstalk signals, improve the separation significantly,

yielding a performance of 30 dB or so at 1 kHz. Improved crosstalk performance is also possible in the envelope decoder by removing the upper frequencies of the S signal by means of an RC filter in the S signal feed to the detector/matrix.

As would be expected, there are hosts of circuits that can be used for stereo decoding, and these are often variations and integrations of the schemes already outlined. In general, however, decoders now almost universally adopt a technique whereby the detector/matrix section performs the dual functions of reinserting the subcarrier in proper phase relationship to the S signal sidebands with demodulation *and* recombining the M and S signals to yield the original A and B signals. This allows a wide range of circuit possibilities, which are extended even more by the three possible arrangements available for producing the 38 kHz subcarrier in the decoder.

DECODER CIRCUITS

As would be expected, decoder circuits differ considerably in complexity. A simple two-transistor circuit based on the time division multiplex technique is given in Fig. 7.10. Here TR1 and TR2 are the decoder transistors proper, while TR3 is used essentially for operating the stereo indicator.

Multiplex signal from the receiver's f.m. detector is fed to the base of TR1 through the filter comprising L1 and C1. A filter like this is often employed to

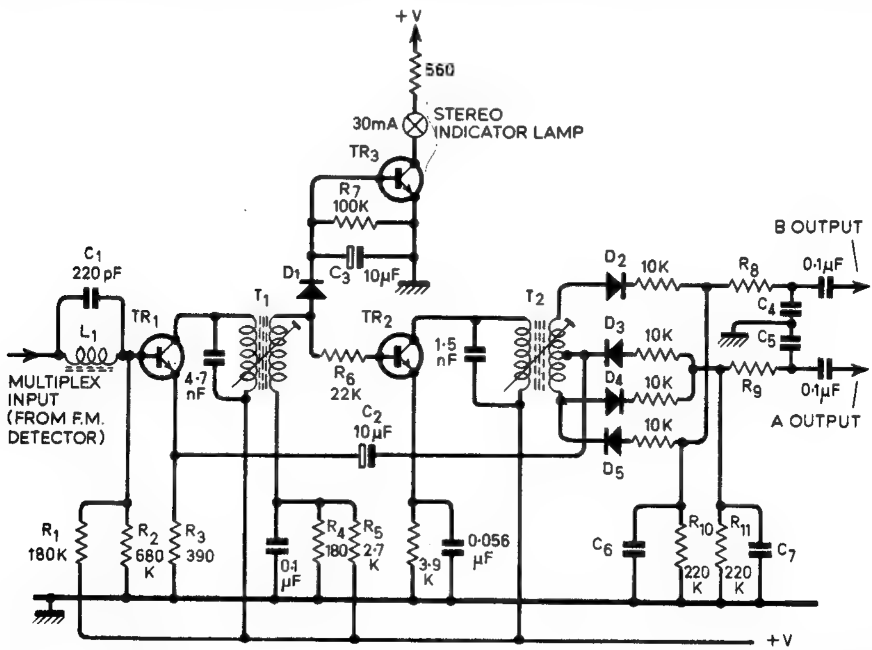


FIG. 7.10. *Simple two-transistor stereo decoder circuit. TR3 is concerned with stereo indicator control, while the filter L1/C1 is for rejecting the SCA subcarrier.*

delete the SCA (Subsidiary Communications Authority) subchannel in areas outside Europe (notably America) in which background music is carried for "piping" into stores and supermarkets. British made equipment aimed for the

American markets might be produced without the filter, therefore, but it would be fitted where the SCA facility is not required.

TR1 is biased by R1 and R2 and T1 in the collector circuit is tuned to the pilot tone frequency (19 kHz). The M and S signals are developed across the emitter resistor R3 and fed through C2 to the matrix transformer T2.

The 19 kHz signal from the secondary of T1 is coupled to the base of TR2, biased by R4 and R5, via R6. This transistor helps with the doubling of the 19 kHz signal to produce the 38 kHz subcarrier signal, as also does diode D1. The load for this diode is R7 with the capacitor C3 across it. Now, transistor TR3 is normally non-conducting (being biased off), but when pilot tone is present at T1 the diode conducts and a positive potential is applied to TR3 base which biases it on, thereby causing collector current to flow through the stereo indicator lamp, resulting in its illumination.

An amplified 38 kHz subcarrier signal thus develops across T2 to drive the balanced ring demodulator composed of the four diodes (D2, D3, D4 and D5) and the 10 k loads. The M and S signals are also applied to the diode ring, at the centre tap of T2 secondary, so when the subcarrier is correctly phased to the S signal sidebands the A and B signals are developed across the loads and fed to their respective outputs through the de-emphasis filters R8/C4 and R9/C5. The time-constants R10/C6 and R11/C7 also provide a degree of filtering and tend to bias the ring demodulator from audio so that the system remains conductive in the absence of pilot tone (e.g., reconstituted 38 kHz subcarrier drive), thereby letting through the audio from a mono transmission.

Fig. 7.11 shows a circuit (Heathkit-Daystrom Ltd) that is based on the envelope detector with switching transistors. Here TR1 receives the complete multiplex signal at its base and delivers it amplified at its collector. The signal is then coupled through the 10 μF capacitor to TR2 base. This transistor is loaded in its collector by T1 and in its emitter by the 560-ohm resistor. T1 is tuned to the pilot tone (19 kHz) while the M and S signals are developed across the emitter resistor for communication to the oscillator/matrix transformer T2.

TR3 in conjunction with the primary of T2 and the associated components forms a 38 kHz oscillator to produce the subcarrier, the signal being locked in phase by the amplified pilot tone fed to TR3 base. The subcarrier appearing across the secondary of T2 relative to the centre tap drives the switching transistors TR4 and TR5 in such a way that the multiplex signal, also applied to T2 secondary centre tap, is sampled alternately on the A and B modulation envelopes (see the waveform at Fig. 7.8a). Here also the matrixing proper is performed in conjunction with the M signal, too, so that the A and B signals appear respectively at the collectors of TR4 and TR5. They are then fed to the left and right audio channels via de-emphasis filters.

So that the switching transistors will conduct in the absence of pilot tone, thereby making the system compatible to mono, they are given a little forward bias by the 15 k resistor to the supply positive line. However, to avoid beat interference effects from the 38 kHz oscillator when the system is processing mono signal, the oscillator is switched off by S2.

The stereo beacon in this example is worked by the two transistors TR6 and TR7. Both are non-conducting in the absence of pilot tone, but when such signal

is present it appears at the base of TR6 from the 200 pF capacitor. This transistor then conducts on the positive peaks of the signal, which causes a rise in potential across the 1 k emitter resistor. This is communicated to TR7 base as a positive "switch-on" potential thus causing the stereo beacon lamp to glow.

In this sort of system with an oscillator generating the subcarrier signal correct phasing is absolutely essential to obtain optimum crosstalk performance. A

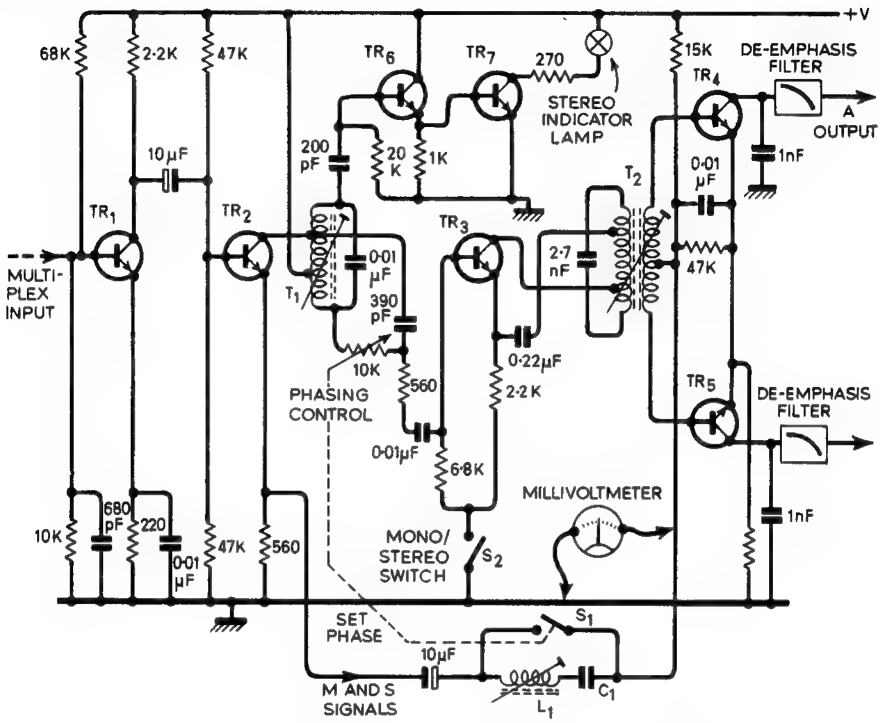


FIG. 7.11. *Circuit diagram of a decoder based on that used in the Heathkit (Daystrom Ltd.) stereo tuners. TR4 and TR5 are switching detectors, driven by the 38 kHz signal from T2. TR3 is a 38 kHz oscillator synchronized by the 19 kHz pilot tone from TR2. TR6 and TR7 operate the stereo indicator.*

clever scheme is here adopted in conjunction with (i) the phasing circuit across the oscillator transformer T1 consisting of the 10 k variable resistor (a potentiometer is used in practice) and the series-connected 390 pF capacitor (it is noteworthy that a 45 deg. phase change of the pilot carrier will reflect a 90 deg. phase change into the reconstituted 38 kHz subcarrier) and (ii) the network in the M and S feed comprising L1, C1 and S1.

Now, when S1 is open L1 and C1 form a filter that deletes all but the 38 kHz subcarrier components of the multiplex signal, and by adjusting the 10 k phasing control the phase of the reconstituted 38 kHz subcarrier can be matched to that of the residual subcarrier components in the signal merely by listening for audio signal in the subcarrier signal. The phasing control is then adjusted until the sound is at a maximum when the filter is in circuit (S1 open). When this condition

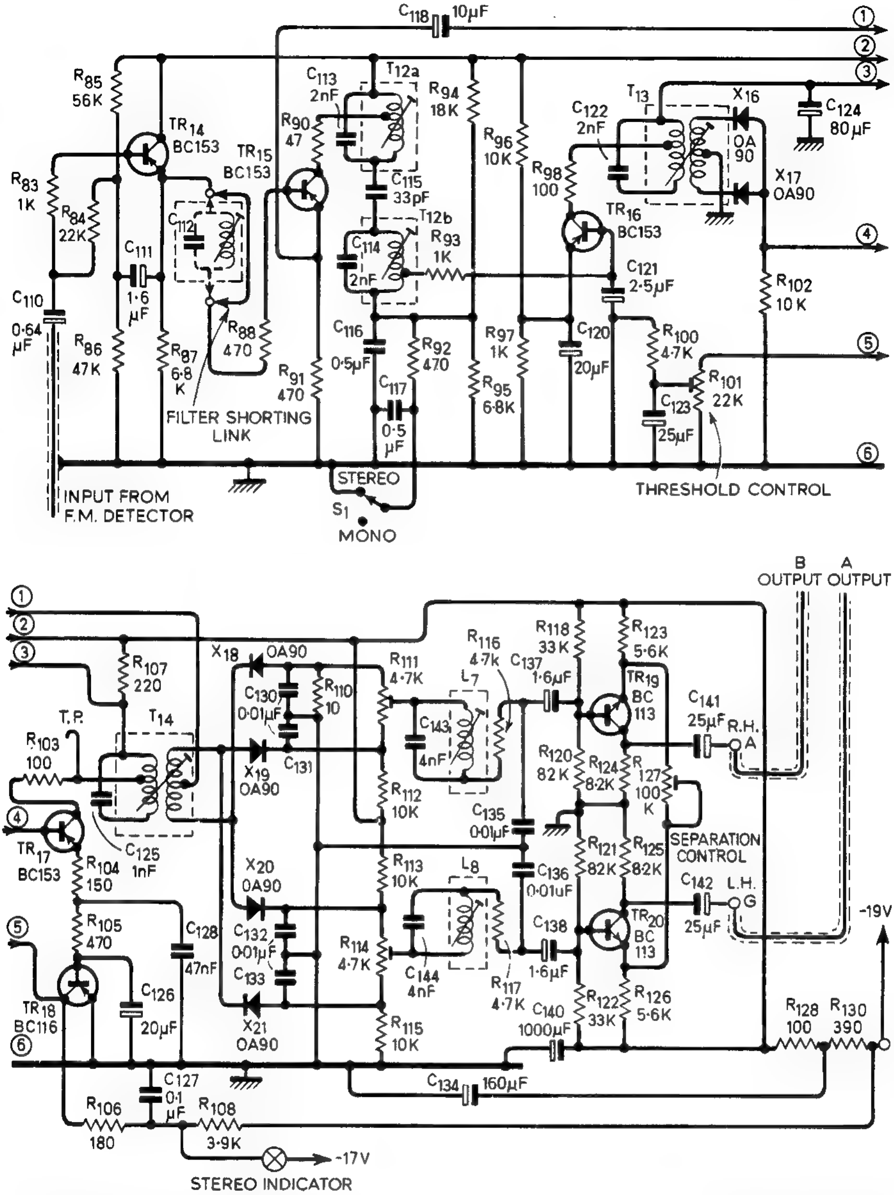


FIG. 7.12. *Circuit of the Goodmans "Stereomax" a.m./f.m. tuner stereo decoder. This is fully described in the text; but note here that the 38 kHz "balance" presets—one in each audio channel—are adjusted in the presence of a multiplex input for minimum 38 kHz components in the audio channels. This model uses one "separation" preset. Some decoders have two, one in each channel.*

has been established S1 is closed, thereby shorting out the filter and restoring normal stereo working.

A circuit of greater complexity (from the Goodmans Stereomax a.m./f.m. Tuner) is given in Fig. 7.12. The first transistor TR14 receives the multiplex

signal from the f.m. detector, and the design of this stage is such that a high impedance (about 350 k) is presented across the detector's load due to the boot-strapping network provided by C111, R84 and R83. This stage is also arranged in common-collector mode, the multiplex signal appearing across the emitter resistor R87.

The signal is then coupled to the base of TR15 via the SCA filter composed of L6 and C112 where applicable. Incidentally, the SCA subchannel is based on 67 kHz, so the filter would be tuned to that frequency. TR15 is wired as a common-emitter to the 19 kHz pilot tone, T12a/b forming a band-pass "transformer" at that frequency which has very good selectivity at 19 kHz, thereby excluding all the other, unwanted at this point, signal components. The M and S signal components are developed across the emitter resistor R91, so to components of those signals the stage acts in common-collector mode. The multiplex signal less the pilot tone is communicated to the centre tap of transformer T14 secondary via C118, referred to again later.

High 19 kHz voltage gain is given by TR15 and the signal, developed at the tap on T12b, is coupled to the base of TR16, which is arranged as a 19 kHz amplifier/limiter. With the stereo/mono switch S1 in the stereo position the base is correctly biased for decoder working from the potential-divider R94/R95, via T12b tap and R93. Emitter stabilization is provided by R96 and R97. The 19 kHz drive fed to the base of TR16 is sufficient to cause signal limiting across T13 forming the collector load, a feature that ensures a constant 19 kHz signal amplitude under all stereo input conditions.

When switch S1 is in the mono position the junction of R92 and C117 is connect to chassis, an action which removes most of TR16 base bias. These two components, in fact, form a filter which prevents the 19 kHz signal from leaving the decoder circuit, where its unwanted presence could cause interference effects. In the mono position, therefore, TR16 is prevented from amplifying or passing the pilot tone, stereo action thereby being destroyed.

Diodes X16 and X17 act as frequency doublers, and thus change the pilot tone, appearing across T13 secondary, into 38 kHz subcarrier signal which, at TR17 base, appears as negative-going pulses having a peak amplitude of about 4·5 V. These pulses are changed into sine waves due to T14 in the collector circuit of TR17 being tuned to 38 kHz also.

Automatic mono/stereo switching is provided by the doubler diodes since TR17 is biased by the d.c. restored doubler output pulses. Thus this transistor is switched on only on stereo, the bias, of course, disappearing on mono due to the absence of the pilot tone.

It will also be seen that TR17 emitter current is passed through the base circuit of TR18, which acts as the stereo indicator switching transistor. The collector of this transistor carries the stereo beacon bulb in series with the supply voltage, so that when TR18 is switched on by the doubled pilot tone, via TR17 emitter circuit, collector current flows and the bulb lights.

It will be observed that in the mono condition the full supply potential exists at TR18 collector (because this transistor is then non-conducting). This potential is applied across the "threshold control" R101. This means that the slider of the control applies a forward bias to diode X15, via R100. Now, under that

condition the base of TR16 is virtually short-circuited signalwise by the low a.c. impedance of the diode in forward conduction and by the 2·5 μF C121. This action makes TR16 insensitive to spurious signals at TR15 collector.

The threshold control is normally adjusted in the presence of a continuous 19 kHz pilot tone until TR18 commences to switch on. This reduces the forward current in X15 and so reduces the signal shunting effect, thereby instigating a kind of chain reaction which results in the circuit switching "solidly" to stereo working. This avoids "hunting" between mono and stereo operation in the event of the pilot tone (e.g., stereo-encoded signal) being of barely sufficient strength for good stereo reception. The circuit will thus be either properly switched to mono or stereo—an intermediate switching state being impossible. Should the stereo beacon bulb fail the threshold circuit action is maintained by virtue of the shunt resistor R108.

Diode pairs X18/19 and X20/21 work rather like the transistors in Fig. 7.11, switching the circuit alternately at 38 kHz between the two multiplex modulation envelopes. It will be recalled that the M and S signals are applied to the centre tap of T14 secondary, via C118, from TR15.

Both the A and B channels work the same, of course, so we need refer only to one channel, say, the B channel. The "mix" of audio, sidebands and 38 kHz components from X18/19 are partly filtered by C130 and C131, while the 38 kHz components and the harmonics of the 19 kHz pilot tone are balanced out by R111. It is noteworthy that the 38 kHz components and the 19 kHz harmonics are of opposite polarity across each diode pair, which is why the balancing presets (R111 under consideration, and R114 in the A channel) cancel these while not affecting the audio components which are of similar polarity.

On a mono signal, when there is no 38 kHz drive to the diode pairs, audio conduction occurs still because of the forward conduction provided by R110 and R112.

The audio signals (mono or stereo) are passed from the sliders of the balance presets to 19 kHz rejectors and de-emphasis filters. The rejector in the B channel is composed of L7 and C143, while the de-emphasis is arranged by R116 and C135.

The B and A audio signals are then coupled to TR19 and TR20 respectively, the transistor pair serving as a differential amplifier to provide crosstalk compensation. It will be seen that preset R127 (the separation control) purposely introduces a small degree of antiphase crosstalk between the channels, the level of which is set by the preset to cancel the slight in-phase crosstalk that results from the process of decoding (see page 125). The audio channels are taken from the collectors while the crosstalk is introduced at the emitters. The voltage gain of the differential amplifier system is about 3 dB, the effect of which just about restores the decoder gain to unity.

CIRCUIT FEATURES

The Goodmans decoder is an excellent example of the various features that are to be found in sophisticated designs which, when geared to a tuner of high sensitivity and low noise, make it possible to receive stereo transmissions outside the normally accepted service areas. Indeed, while writing this chapter I am

receiving really good stereo reception from a 22 kW (e.r.p.) relay station some 160 miles from my home! I must admit, though, that my altitude is about 180 ft above sea level and the aerial is a six-element array specially designed for long distance f.m. reception by J-Beam Aerials Ltd; but it is in the roof-space.

Another automatic mono/stereo switching circuit, by Lafayette (in Model LR500T tuner-amplifier), is given in Fig. 7.13. Here D1 and D2 are the doubling

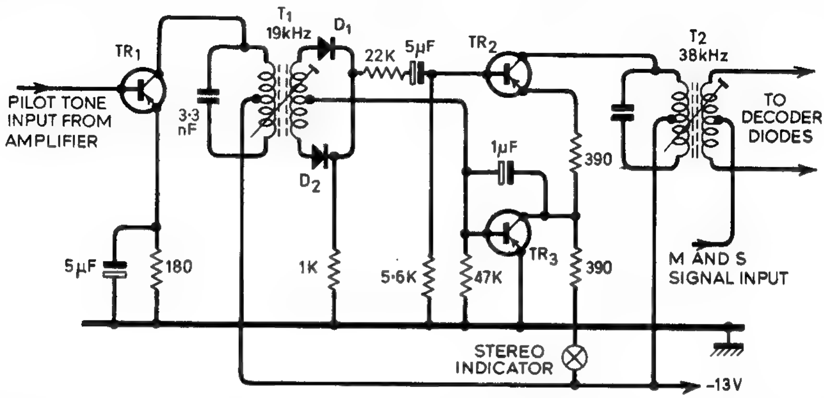


FIG. 7.13. *Auto mono/stereo switching circuit used in the Lafayette LR500T tuner-amplifier. The circuit also operates the stereo indicator and, to some degree, works in conjunction with the circuit in Fig. 7.14 (see text).*

diodes fed from pilot tone via amplifier TR1 and transformer T1. The 38 kHz signal is coupled to the base of TR2, via the 5 µF capacitor, and when this transistor is conducting the 38 kHz subcarrier is developed across T2, which drives the decoder diodes and which also receives the M and S signals at its secondary centre tap.

Now, 19 kHz pilot zone is also fed to the base of TR3 from a tapping on T1 secondary which causes this transistor to conduct, thereby putting TR2 cathode to chassis, via the 390-ohm resistor, and causing that transistor to conduct also, so passing subcarrier signal to T2. When TR3 conducts the collector receives a current via the stereo indicator bulb and the other 390-ohm resistor. The bulb thus lights indicating that a stereo transmission is being received. The circuit can only operate, therefore, in the presence of a pilot tone.

A manual mono/stereo switch is also provided, so that under conditions of a weak stereo transmission the auto circuit can be overridden to prevent hunting between the stereo and mono modes. The circuit is given in Fig. 7.14, where TR1 is a pilot tone amplifier which, in the stereo mode set by switch S1, is biased by a negative potential derived from a signal rectifier located towards the end of the tuner's i.f. channel. The biasing is through the 5·6 k and 12 k resistors connected to S1, with the 12 k threshold preset forming a potential-divider leg. Thus, the level of bias at which TR1 is caused to conduct is set by the preset. The i.f. signal —and hence the stereo transmission—thus has to be of a certain, preset strength, before TR1 will conduct and allow the circuit to operate in the stereo mode. With S1 in the mono position the base circuit of TR1 is connected direct to chassis and the transistor remains non-conducting. The signal rectifier just

mentioned also operates the tuning meter, the tuning being adjusted for maximum meter deflection.

Some hi-fi receivers and stereo tuners have a position on the selector switch labelled "multiplex filter" or similar. In its simplest form this merely causes an

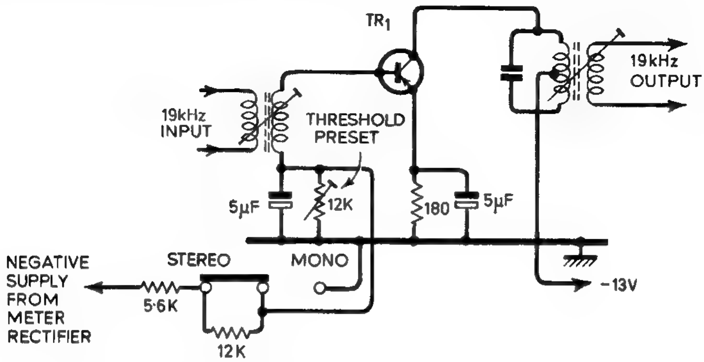


FIG. 7.14. *Stereo/mono manual switch in the Lafayette LR500T.*

extra roll-off at the treble audio frequencies. The normal de-emphasis filters, of course, yield a 6 dB/octave roll-off at a turnover frequency of 3·5 kHz (response —3 dB at that frequency), but for extra noise immunity, with some slight loss in treble, depending on the nature of the filter. In fact, it is surprising how much assistance in this connection is given by the addition of a capacitor across the audio outputs of a decoder, and it is not uncommon for a capacitor—about 2·2 nF—to be shunted across these circuits in the "filter on" position. This sort of filtering rids the stereo channels of excessive high-frequency "sizzle" which can result when the stereo signal is not all that strong at the aerial.

FILTERS FOR TAPE RECORDING

Some of the less complex decoders have only basic filtering of the 19 kHz pilot tone and harmonics of this and of the 38 kHz subcarrier signal at the audio outputs, so when a decoder is used to feed a stereo signal into a tape recorder "beat interference" between these spurious signals and the recorder's bias oscillator not uncommonly results in a whistle (or "birdies") accompanying the recording—being actually recorded with the audio signal on to the tape.

The solution to this problem lies in more efficient filter at the A and B signal outlets, but this has to be done in such a way as to avoid the treble roll-off starting too early and impairing the treble performance of the recording or system generally. A simple RC filter is not good enough for this application, therefore.

It is possible to employ tuned filters (e.g., rejectors) resonating at 19 kHz and 38 kHz, but these are not much good for harmonics and their inter-beats. The most effective means of overcoming these sorts of beat troubles is by the use of an active feedback filter, an example of which is shown in Fig. 7.15. Here TR1 and TR2 are in d.c. complementary mode, with feedback from TR2 collector being to TR1 base, the feedback frequencies controlled by the RC elements, with tailoring of the response made possible by the 5 k preset. Rejection is about —15 dB at 19 kHz relative to 1 kHz, and by adjusting the preset it is possible to

move the turnover frequency along the response curve; but for the above figures it is adjusted so that the response at 10 kHz is the same as at 1 kHz.

Parallel-T "notch filters", both passive and active, are also useful for notching out the 19 kHz and 38 kHz components, and at least one firm is marketing active filters of this nature embodying an emitter-follower circuit to give a high input

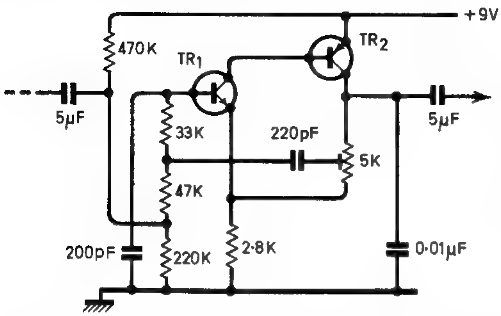


FIG. 7.15. *Active filter for deleting "birdies" and "warbles" resulting from multiplex components "beating" with the bias oscillator of a tape recorder. The transistors should be low noise audio types.*

impedance and a low output impedance, parameters which facilitate matching between almost any decoder and tape recorder. Of course, for stereo recording two such filters are required, one in each channel.

Some f.m. tuners and hi-fi receivers (e.g., tuner-amplifiers) are designed with the stereo decoder already fitted; different models, on the other hand, are designed in such a way that a stereo decoder can be easily plugged in when the equipment is operated in an area served by a stereo transmitter. At one time stereo "converters" were popular, too; however, the trend is now towards stereo-ready tuners and receivers, irrespective of the reception area. As already intimated, the improvements in tuner sensitivity and noise performance, brought about the use of high-gain integrated circuits and field-effect transistors in the first stages, are making it possible to receive stereo programmes in areas not so long ago considered far too remote from a stereo-encoded station; moreover, as time goes on more transmitters are being updated for stereo emission.

Fig. 7.16 shows a stereo decoder "module" fitted in one make of tuner, while Fig. 7.17 shows an early stereo "adaptor" by Heathkit (Daystrom Ltd). An inside view of the Goodmans f.m./a.m. stereo tuner is depicted in Fig. 7.18.

Some interesting BBC pictures concerning the stereo system are shown in Figs. 7.19, 7.20, 7.21 and 7.22.

SERVICING DECODERS

If a stereo tuner or receiver operates normally on mono, but fails to work properly on stereo, one can be sure that the audio channels and the M signal input to the decoder diodes are active. As it is very difficult to tell merely by listening to a stereo programme whether or not the decoder is working correctly, the BBC radiate a series of stereo test signals following the conclusion of programmes on Radio 3. Every day except Wednesday and Saturday 250 Hz tone is transmitted in the left channel only from about four minutes after the end of Radio 3 until 23.55 hours (subject to possible interruptions from time to time as may be demanded by transmitter testing etc.). This test makes it possible (i) to identify the A and B channels and (ii) to check on crosstalk at 250 Hz.

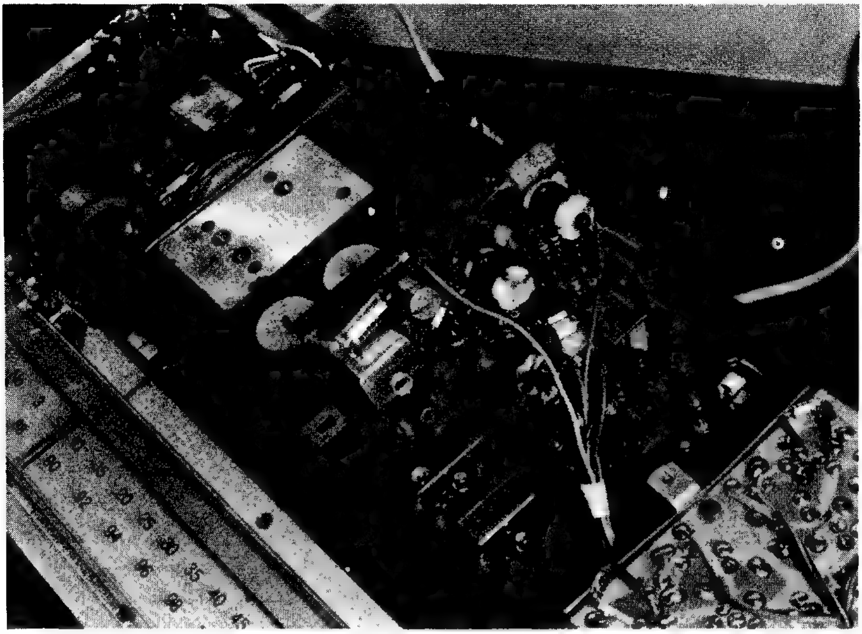


FIG. 7.16. *Stereo decoder "module" used in one f.m. stereo tuner, the Bang and Olufsen Model 1000.*

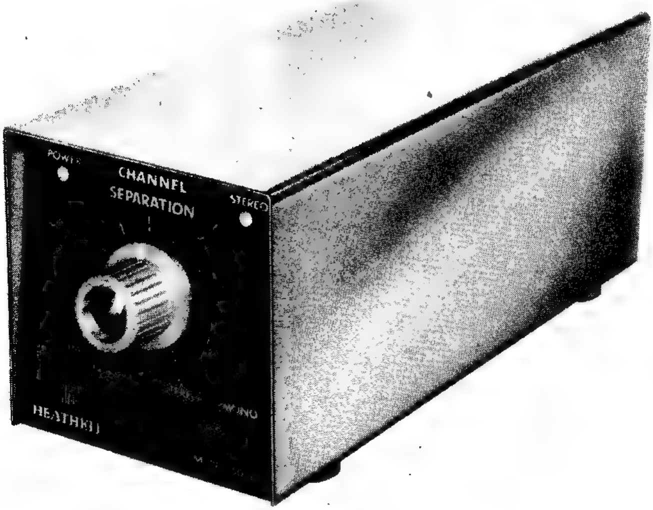


FIG. 7.17. *"Stereo Adaptor" by Heathkit (Daystrom Ltd.) for updating a mono f.m. tuner. The circuit contained in this "adaptor" is utilized in some of the firm's stereo tuners and tuner/amplifiers.*

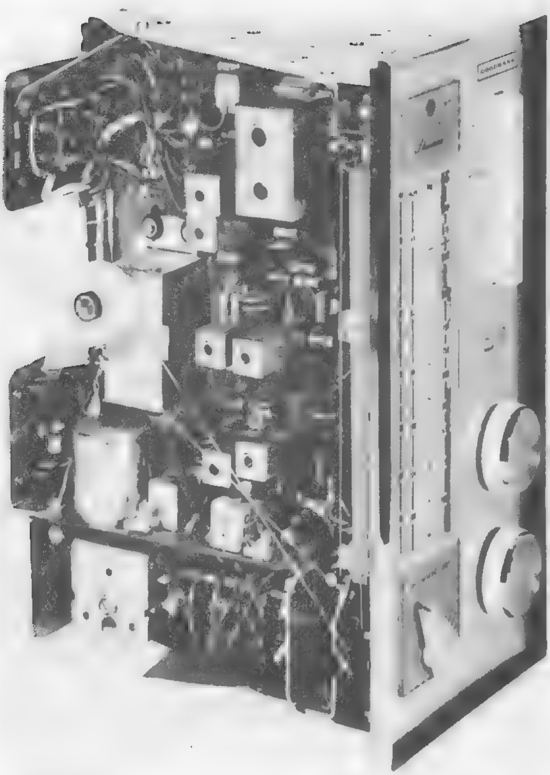


FIG. 7.18. *Inside view of the Goodmans "Stereomax" a.m./f.m. tuner.*

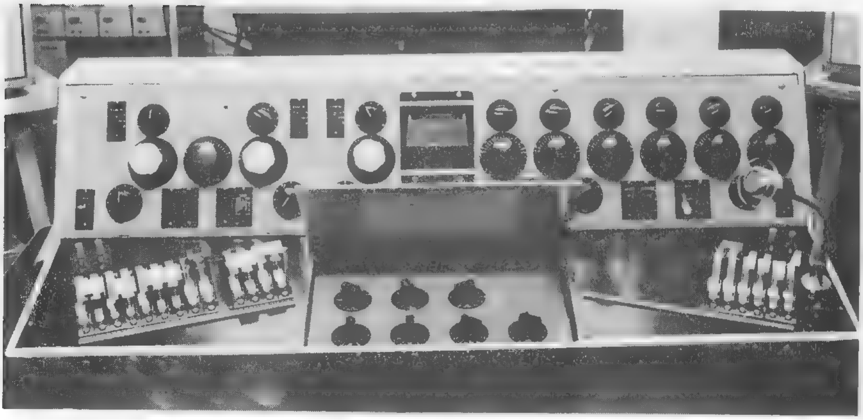


FIG. 7.19. *BBC's transportable stereo control desk with flap raised to reveal presets (BBC Photo).*

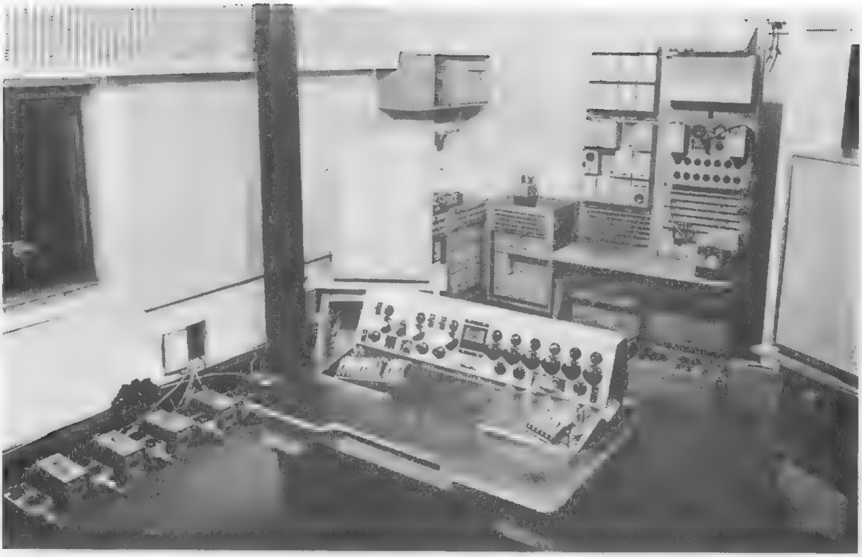


FIG. 7.20. *BBC's stereo equipment installed in the Playhouse, Manchester (BBC Photo).*

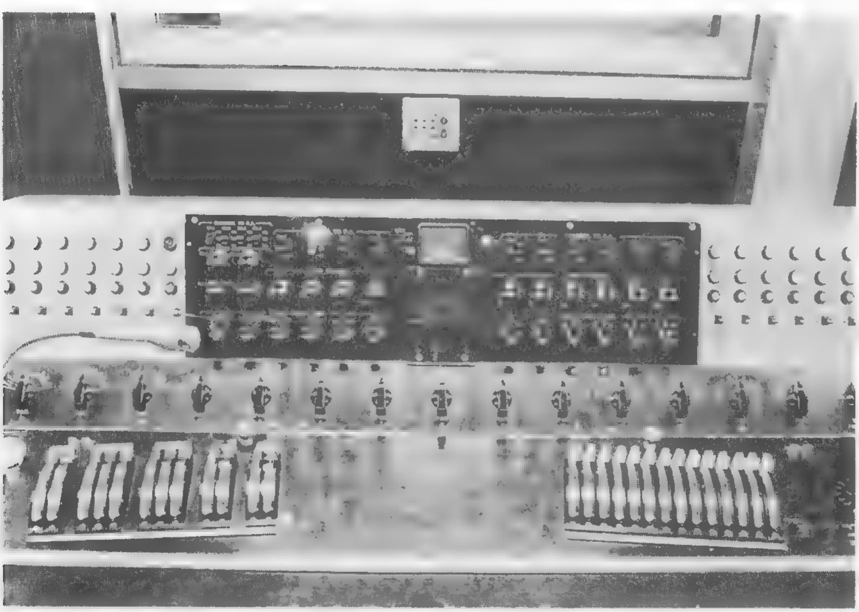


FIG. 7.21. *Close-up of BBC's transistorized dual-purpose mono/stereo studio equipment (BBC Photo).*

FIG. 7.22. *In the stereo studio of the BBC. This is the control room of Studio 10 in which The Archers was being recorded (BBC Photo).*

A more extensive test sequence is radiated on Wednesdays and Saturdays, starting at 23.30 hours from all stereo transmitters, the schedule of which is shown in Table 7.1.

Table 7.1

Test No.	*Time*	*A Channel*	*B Channel*
1	23.30	250 Hz at zero level*	440 Hz at zero level
2	23.32	440 Hz at zero level	440 Hz at zero level (antiphase to A channel)
3	23.35	440 Hz at +8 dB	440 Hz at +8 dB (antiphase to A channel)
4	23.37	440 Hz at +8 dB	440 Hz at +8 dB (in phase with A channel)
5	23.39	250 Hz at +8 dB	440 Hz at +8 dB
6	23.40	250 Hz at zero level	Nothing
7	23·44	Nothing	440 Hz at zero level
8	23.47 +20 s**	Tone sequence at −4 dB: 60, 900 Hz and 5 and 10 kHz; sequence repeated	Nothing
9	23.48 +20 s**	Nothing	Tone sequences as for Test No. 8, A channel
10	23.49 +20 s	250 Hz at zero level	Nothing
11	23.51	Nothing	Nothing

Notes: * Zero level corresponds to 40 per cent of the maximum level of modulation applied to either stereo channel before pre-emphasis. All tests are transmitted with pre-emphasis.
** approximately.

The schedule is subject to variation to accord with programme requirements and essential transmission tests, and the periods of tone lasting several minutes are interrupted momentarily at one-minute intervals.

USING THE TEST TONES OF THE BBC

Let us now see how this test sequence can be used for checking and adjusting a stereo decoder. Test No. 1, of course, is fairly obvious, it being necessary merely to listen for 250 Hz tone in the left speaker and 440 Hz tone in the right speaker. If the tones are reversed, and the stereo tuner is connected to a main amplifier system through a pair of screened cables, one would then normally change over the cable connections. However, it could be that the loudspeakers are connected in reverse at the amplifier, in which case the stereo placement on the other programme sources, such as gramophone record and tape, would also be the wrong way round.

Test No. 2 is provided to allow fairly simple adjustment of the phase of the subcarrier. It must be assumed to start with, though, that the 19 kHz and 38 kHz tuned circuits have already been adjusted for maximum subcarrier output, and it is not always easy to tell whether this is the case or not from the test transmission alone. However, if the decoder incorporates a stereo indicator, the 19 kHz tuned circuit can generally be peaked to produce maximum illumination from this bulb, and to secure a really accurate adjustment on this circuit it is desirable to attenuate the aerial signal (a series of plug-in Belling and Lee attenuators are useful for this or, better, a switched 75-ohm v.h.f. attenuator) progressively after first roughly tuning on a strong signal. The two 19 kHz tuned circuits in Fig. 7.14 could be adjusted like this, as also could the tuned transformer T1 in Fig. 7.13, since the stereo indicator is actuated from the regenerated subcarrier *after* T1. T1 in Fig. 7.10 is also before the stereo indicator transistor TR3. In Fig. 7.12 both T12 and T13 could be tuned to 19 kHz using the stereo indicator; but here, and in circuits like this one, the limiting action of the pilot tone and subcarrier transistors would make it virtually impossible to tune for maximum illumination, for which decoders of this kind, with anti-hunting devices, a certain level of signal is required for switching to stereo, the switching then almost immediately building up to the "full on" condition.

It is also possible to apply a 19 kHz signal from an audio generator to the decoder input and then monitor the signal on an oscilloscope or an electronic millivoltmeter in terms of 38 kHz drive to the decoder diodes or switching transistors, depending on the nature of the circuit. The idea, then, of course, is to adjust all the tuned circuits for maximum amplitude as indicated on the oscilloscope or millivoltmeter. The transmitted pilot tone can also be used in conjunction with such a monitoring device, one then knowing for sure that the input is truly at 19 kHz.

Test No. 2 provides for a "trimming" adjustment, so to speak, and since the 440 Hz signal is antiphase to the left channel during this test, the tuned circuits should be very critically adjusted, slightly one way or the other relative to the peaked setting previously established, until the output is *maximum* in the "speaking" channel. It must again be mentioned that a 45 deg. phase displacement of the 19 kHz pilot tone at the decoder will result in double the phase displacement

at the 38 kHz subcarrier frequency—e.g., 90 deg. It is sometimes noticed that two apparently correct phasing points can be achieved when adjusting the 19 kHz tuning. The correct one is that which yields the best separation and noise performance; the incorrect one might reverse the channels!

Test No. 3 is for checking distortion with the signal wholly in the S channel (e.g., A — B channel), while Test No. 4 is also for checking distortion but this time wholly in the M channel (e.g., A +B channel). These tests are achieved by the phasing being switched at the transmitter. A distortion check with the signal divided equally between the M and S channels is provided by Test No. 5. To secure an accurate assessment of the distortion, of course, a distortion test set tuned to 440 Hz or a steep notch filter at that frequency is required along with a very sensitive millivoltmeter to give the percentage of harmonics relative to the full amplitude of the audio sine wave at the decoder output.

A to B and B to A crosstalk checks are provided by Tests 6 and 7 respectively. There are some decoders in which the only control of crosstalk lies in the adjustment of the subcarrier phase (or pilot tone tuning), and this is best made on Tests 6 and 7 rather than on Test 2, after first having established the near-correct phasing as already explained. When a separation control or preset is featured (two presets sometimes), as in Fig. 7.12 circuit (also see Fig. 7.23), adjustment

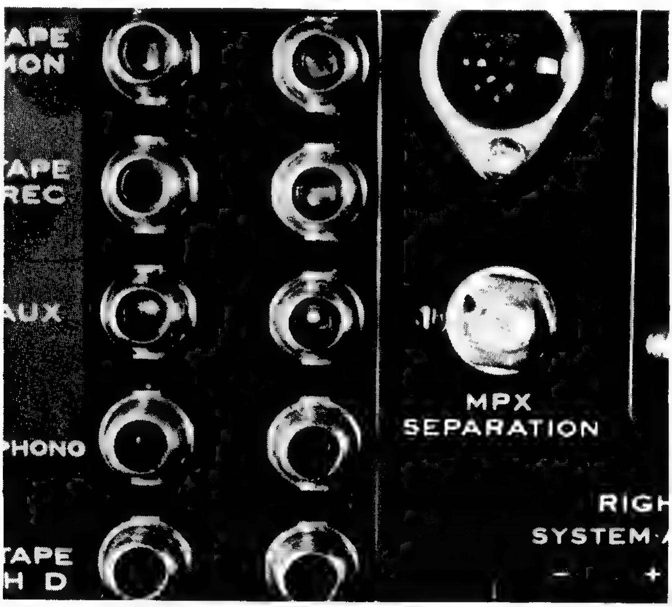


FIG. 7.23. *Multiplex "separation" preset as often found at the rear of a stereo tuner or receiver. Sometimes there are two presets, one for each channel.*

should be made for the least output in the "non-speaking" channel. This calls for a compromise adjustment between the two channels obtained, if possible, during the course of several A and B changeovers (but this would demand one's own f.m. stereo generator). The optimum value results from the balancing-out of asymmetries in the decoder: it is noteworthy that the separation control rarely provides a null or minimum adjustment related to both channels, merely

an optimum setting, established by compromising a little between the two channels.

Incidentally, a better crosstalk performance can sometimes be achieved by unbalancing the separation control (turning it to range extreme or, preferably, disconnecting one end) and then adopting the subcarrier phasing adjustment for the very best crosstalk performance, after which reconnecting the separation control and adjusting to delete as much crosstalk as possible from the "non-speaking" channel. When adjusting the crosstalk by the phasing method, it is rarely possible to obtain zero output on the "non-speaking" channel, and harmonic components of the 440 Hz modulation are usually detected. These, of course, would normally be attenuated by the action of the separation control, since at the decoder output the phases of the crosstalk components are in opposition, as mentioned in connection with the Goodmans decoder.

Subcarrier filtering is sometimes provided by an RC parallel combination between the M and S output of the first amplifier (usually at the emitter of a transistor as we have already seen) and the centre-tap of the detector drive

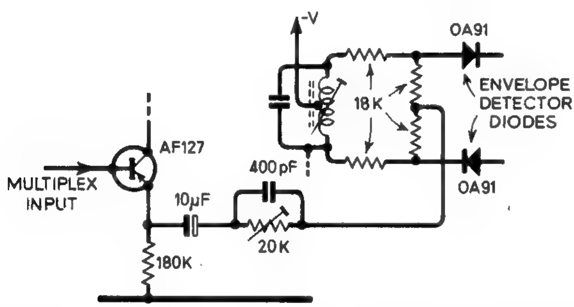


FIG. 7.24. *Filtering is sometimes adjustable as shown here with R being the variable element of an RC network.*

transformer secondary or equivalent circuit. An arrangement in which two diodes in an evelope detector are used is shown in Fig. 7.24.

Regarding frequency-doubling of the pilot tone to form the 38 kHz sub-carrier, it should be noted that this process incites a 90 deg. phase lag on the regenerated subcarrier, and to compensate this the pilot tone is given a 45 deg. phase shift, often by means of an RC network at a vantage point in the circuit. As already mentioned, a 45 deg. shift at 19 kHz is reflected as a 90 deg. shift at 38 kHz.

So much, then, for the main adjustments which can be handled on the test transmission or by an a.m. generator with stereo-encoding built in. Hi-fi shops, who will be handling more and more stereo tuners as time goes on, usually have access to a stereo generator, since it would be frustrating, to say the least, to have to wait until almost midnight to use the BBC's tests. The enthusiast, though, is less dictated to by time in this respect. The BBC's tests are referred to again at the end of this chapter.

An oscilloscope, too, is a fundamental item of test equipment required for successful decoder servicing. Such an instrument makes it easily possible to trace

the various signals from the f.m. detector all the way through the decoder to the A and B outputs. My book entitled *Servicing With The Oscilloscope*, by the same publisher, gives information on the use of the oscilloscope in decoder servicing, along with off-screen waveform displays (oscillograms).

The decoder embodying a 19 kHz or 38 kHz oscillator can only be checked adequately subcarrierwise by means of an oscilloscope. One way of aligning this sort of decoder, such as that in Fig. 7.11, is by means of an audio generator and millivoltmeter, applied as follows relative to the circuit in Fig. 7.11.

Disconnect the input of the decoder from the f.m. detector and apply a 38 kHz signal at a level of about 10 mV to the base of TR1, via a 10 μF capacitor. Connect the millivoltmeter between chassis (earth) and the M and S input to the switching detectors, as shown in the circuit. Arrange the "set phase" control so that S1 is "open" and adjust L1 for maximum reading on the millivoltmeter. This will be about 200 mV. Change the input to 67 kHz at 100 mV and close S1 with the phasing control about two-thirds advanced. The SCA filter, if fitted, should then be adjusted for minimum reading on the millivoltmeter. This is often located, as in the Heathkit, between the f.m. detector and the decoder input; though it has a slightly different location in the Goodmans circuit in Fig. 7.12.

The next move is to connect the decoder back to the f.m. detector and, with S1 open, adjust T2 on a stereo transmission, preferably the test tones, for a clear output by adjusting the core one way and then the other until the correct setting is found. Next the critical core point where the sound becomes "garbled" should be established. From that setting the core should be advanced by a quarter turn and T1 adjusted (this is the 19 kHz oscillator transformer) until the minimum sound (null point) is obtained. When this is correct the stereo indicator will have maximum illumination. The phase control is then turned anticlockwise until the point of maximum sound output is obtained, after which S1 should be closed to give the correct stereo working.

Different circuits and designs will, of course, have different setting-up procedures, but those detailed above will be suitable for the majority of decoders; though it is always best to follow the maker's instructions where these happen to be available.

If there happens to be any particular fault in the decoder this is almost always brought to light in terms of lack of response of one or more of the tuned circuits when adjustments are attempted to produce the best separation and, possibly, when checking for crosstalk performance. An oscilloscope connected to vantage points in the decoder will also reveal quickly the part of the circuit failing to pass signal or process it properly, it then remains merely to test the suspect area by adopting ordinary servicing techniques. If a decoder fault is responsible for even the M signal failing to pass to the decoder detectors, this can be checked pretty quickly by coupling the signal at the f.m. detector output of the tuner proper direct to the A or B audio channel, thereby bypassing the decoder. Remember, though, that this action will possibly delete the normal de-emphasis and cause the reproduced audio to have a significantly rising treble characteristic. If there is an output, however, one can be sure that the trouble lies in the decoder and not in the tuner itself.

REQUIREMENTS FOR GOOD STEREO

Even though the decoder is correctly fitted and aligned properly, good stereo performance is possible only if the tuner circuits themselves are compatible with this 'ultra' wideband multiplex signal. With ordinary mono f.m. reception the modulation frequencies extend little beyond the 15 kHz mark (at the present, anyway); but with stereo multiplex we have S signal components going well up to 53 kHz. This means that more critical i.f. parameters are called for with a reasonably wide i.f. pass-band and linear f.m. detector characteristics. Earlier chapters revealed that the nature of the f.m. system produces series of pairs of sidebands, so it might be expected that an i.f. pass-band greater than that required for mono would be needed for stereo. This is not true, however, for the energy contined in the higher-order sidebands is very small and all the useful energy can be handled in the fairly normal pass-band of about 180 kHz provided this is achieved with an essentially linear phase response.

It is noteworthy, though, that some mono f.m. tuners, for reasons of enhanced selectivity and sensitivity, run with a pass-band of about 150 kHz (sometimes, sadly, even less, which is really below the mono minimum). A multiplex signal running through such channels would tend to suffer undue attenuation of the subcarrier and S-signal components, and good stereo performance would be nigh impossible.

While it was at one time considered desirable from the "quality" aspect to employ a phase discriminator rather than a ratio detector in hi-fi tuners and receivers, the trend is now towards the ratio detector in all grades of receiver and tuner, mono *and* stereo, using the balanced circuit. Indeed, since the advent of stereo it has been discovered that remarkably good performance is possible by means of a well designed and balanced ratio detector. However, the output circuit or i.f. filtering network of the ratio detector requires an a.f. bandwidth of 53 kHz when engaged with stereo, while only 15 kHz is needed for mono. Thus, when fitting a stereo decoder to a mono tuner special pains must be taken to ensure that the a.f. filter is regeared and the de-emphasis removed from the f.m. detector output circuit. The a.f. filtering must, as just mentioned, have a bandwidth of at least 53 kHz.

We have seen that the depth of modulation remains the same with stereo as with mono, but while some non-linear working (towards the edges) can be tolerated with mono, such is highly undesirable with stereo for several reasons. One is that non-linearity can lead to distortion and phase upsets which can impair the stereo separation; another is that beats between the various multiplex components can be intermodulated, so to speak, due to the poor linearity towards full modulation peaks; and a further reason, and one which is just about coming into full significance at the time of writing, is that the multiplex components arriving at the f.m. detector when running into non-linearity from a wanted stereo-encoded station and from an adjacent channel, unwanted stereo-encoded transmission can produce highly undesirable beat-interference effects, manifesting as a "warbling" sound or "birdies" in the background of the reproduced, wanted station.

WARBLING EFFECTS

We have seen that channels are separated by 200 kHz in the f.m. band; but many sets and tuners are hard-pushed in discriminating adequately between two fairly powerful stereo aerial signals spaced by a mere channel width, especially when the i.f. amplifier is tailored for a bandwidth of, say, a minimum of 180 kHz with low phase distortion. On mono transmissions in adjacent channels the capture effect will, of course, pick out the strongest one—that to which the set is accurately tuned—and push the off-channel unwanted one well into the background, but when the transmissions are stereo-encoded the "capture effect" laws fail to apply in quite the same way on the multiplex part of the signals. Hence the warbling or bird noises.

It is desirable, therefore, for the f.m. detector to have a bandwidth in excess of that of the i.f. channel; and very accurate alignment and symmetry are

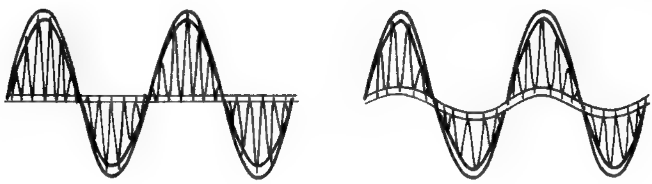


FIG. 7.25. *An f.m. receiver whose output from a stereo signal modulated in the A channel only is as shown at (a) would be suitable for driving a stereo decoder. If the signal is at (b), with a wavy axis, then the receiver might not be suitable unless the waviness can be eliminated by realignment. The greater the waviness, the less suitable is the receiver for stereo working.*

essential. These can only be achieved by visual alignment, using the wobbulator and oscilloscope partnership, explained in Chapter 9.

Even when all precautions have been taken to avoid the beat effects mentioned, very slight background warbling might be experienced in areas where two stereo-encoded transmissions can be received on adjacent channels on the one aerial in common orientation, and when the signals presented to the tuner or receiver are fairly strong. Reports have also been received of similar symptoms when two such stations are spaced by two f.m. channels (e.g., 400 kHz). In this extreme case, of course, the tuner selectivity, i.f. response characteristics and f.m. detector characteristics would be suspect. When the unwanted interfering transmission is inducing a high value of signal voltage into the aerial, attempts should be made to secure better discrimination against this in favour of the wanted transmission by re-orientating the aerial or, if necessary, employing a more directional array to yield the required discrimination. Sadly, this is not valid when the unwanted transmission happens to be arriving on a path in line with the aerial when it is, in fact, "beamed" towards the wanted transmission, anyway! There are some areas of the country where this happens.

Designers of stereo receivers and hi-fi tuners have this problem in hand at the time of writing, and one major requisite, so it would seem, is for a low-pass filter between the f.m. detector output and the decoder input which, while passing all components up to 53 kHz, has a fairly fast roll-off at frequencies beyond this,

the design being to avoid major phase shift in the spectrum occupied by the multiplex components. Such a filter deletes the high order harmonics of the multiplex components which appear to be responsible for the "warbling" and "birdies".

Returning now to the i.f. bandwidth aspect, it is not always desirable to attempt to improve the stereo performance by readjusting the i.f. tuning (even visually) for a wider bandwidth when the design of the i.f. stages is geared to a mono bandwidth of, say, 150 kHz. The reason for this is that the increased bandwidth could be negated by a resulting impairment in phase response. Thus, increasing the bandwidth without giving sufficient thought to the phase parameter could well make matters worse rather than better!

MORE ABOUT THE TEST TONES

The nature of the multiplex waveform appearing at the output of the f.m. detector without de-emphasis can be used as a guide to the receiver's suitability for stereo working. The stereo signal transmitted by the BBC with 250 Hz tone in the A channel only is ideal for this check, and Fig. 7.25 shows at (a) a good stereo waveform. However, if the signal is displayed as at (b), then the receiver is either misaligned or unsuitable for working a stereo decoder properly. These tests should be performed with the decoder disconnected from the f.m. detector.

When a decoder is running properly the separation over the audio spectrum should not fall much below 20 dB, and this can be checked on BBC Tests 8 and 9 for A on B and B on A respectively. The separation should be at least 20 dB at 440 Hz on Tests 6 and 7. Many decoders linked to well designed and adjusted tuners yield separation figures better than 30 dB at 440 Hz.

Tests No. 10, giving 250 Hz in the A channel only, provides a reference level for Test No. 11, which is for noise, the modulation being cut from both channels.

CHECKING CROSSTALK AND S/N RATIO

A sensitive audio voltmeter or millivoltmeter scaled in decibels is required for checking both crosstalk and noise. For crosstalk a 0 dB reference should first be established on the "speaking" channel from the maximum available audio (e.g., full limited output of the tuner or receiver) at the A or B output. When the channel goes "non-speaking" the sensitivity of the meter should be increased either by switching out attenuation or switching to lower ranges until the original 0 dB reference is again obtained. The amount of attenuation switched out or the amount of sensitivity increase of the meter, in terms of decibels, is a direct, unweighted measure of crosstalk in that particular channel. The process should be repeated in the other channel.

For noise, the 0 dB reference should be established on Test No. 10, and the meter sensitivity increased or attenuation reduced until the same 0 dB reference is obtained on Test No. 11. Both A and B channels can then be checked (since both are "non-speaking" on Test 11) for noise ratio in turn. For hi-fi listening the signal/noise ratio should be at least 50 dB; but a ratio in this order—or better—demands a fairly strong aerial signal and a receiver with a low-noise front-end. The action of the regenerated 38 kHz subcarrier in switching the decoder detectors and the greater a.f. bandwidth add significantly to the noise

in the stereo mode (see aerial requirements in Chapter 8).

Table 7.2 should assist with fault diagnosing in stereo receivers and decoders.

Table 7.2

Symptom	*Probable cause*	*Check*
No sound	Faulty receiver	Audio at f.m. detector
	Lack of audio continuity in decoder	Decoder input amplifier. M and S feeds to multiplex detectors. Biasing of these detectors. Decoder transistors and voltages
Weak sound	Faulty receiver	Audio at f.m. detector
	Signal attenuation in decoder	Decoder transistors and voltages (biasing)
	Misalignment	Alignment
Distortion	Misaligned receiver	Receiver alignment
	Misaligned decoder	Decoder alignment
	Incorrect bias on decoder transistor	Transistor voltages and biasing
	Faulty decoder transistor	Decoder transistors
	Faulty decoder switching detector	Switching diodes or transistors
	Decoder detector low forward bias (on mono)	Forward biasing of detector diodes or transistors
Hum	Poor h.t. smoothing	Smoothing capacitors
	Incorrect audio couplings	Screened circuits and shielding
High noise	Low aerial signal	Aerial (see Chapter 8)
	Faulty receiver	Audio for noise at f.m. detector
	Noisy transistor or component in decoder	Appropriate components
	Defective decoder tuned circuit	Tuned circuits (L *and* C)
No stereo (beacon not lit)	Pilot or subcarrier stage defective	Appropriate area
No stereo (indicator lit)	Subcarrier circuit faulty	Doubler stage, oscillator and decoder detectors
	Misaligned subcarrier tuned circuits	Decoder alignment
	Bad aerial	Aerial
Poor separation (indicator lit)	Decoder and/or tuner misaligned	Alignment
	Incorrect subcarrier phasing	Phasing (see text)
	Incorrectly set crosstalk preset or presets	Setting of controls
	Incorrect aerial or misorientation Multipath interference	Aerial, signal strength and orientation
Good stereo but indicator out	Beacon control circuits Beacon lamp blown	Transistors and components Beacon lamp

Symptom	*Probable cause*	*Check*
Hunting between mono and stereo	Auto control circuit faulty	Auto control circuit and semi-conductors
Poor stereo (indicator flashing on and off)	Aerial signal poor	Aerial etc.
Warbling or "birdies" on stereo	Poor tuner selectivity	Tuner alignment
	I.F. misalignment	Tuner alignment
	Poor f.m. detector response (non-linearity at extremes)	F.M. detector tuning
	Too strong adjacent channel (stereo-encoded)	Reorientate aerial. Try more directional aerial
	Insufficient low-pass filtering between f.m. detector and decoder input	Try fitting a filter rolling off at 60 kHz
	Tuner design fault	Complete realignment *might* help
"Birdies" when replaying tape recorded stereo programme	Harmonics of 19 kHz/38 kHz multiplex beating with recorder bias oscillator	Fit suitable filters (see text)

CHAPTER 8

V.H.F. and F.M. Aerials

IT WAS not all that long ago when it was thought that the v.h.f. spectrum held little value for domestic broadcasting because of the fact that v.h.f. signals cannot normally be received adequately over long distances. Of recent years, however, it has been demonstrated that the spectrum is extremely useful for relatively short-range (local) propagation of high-definition monochrome and colour television and monophonic and stereophonic f.m. sound-only transmissions.

The factor of limited range of v.h.f. transmissions has now been accepted and, although intially—before television and f.m. were seriously contemplated as domestic services—it may have been considered as an adverse factor, it is now found to possess a desirable influence, inasmuch as it ensures the minimum of interference from relatively distant stations which may be operating in the same channel to which a receiver may be tuned to receive the local programme. Briefly, the factor permits controlled channel-sharing without the product of interference.

As opposed to the common practice of calibrating the tuning scales of ordinary broadcast receivers in wavelength, the tuning scales corresponding to Band II on f.m. receivers and f.m. tuners are calibrated in frequency. The reason for this is that the wavelength conversion would be very cumbersome. It is easier to read 88 MHz on the scale, for example, than to read the wavelength conversion, which is 3·40909 metres. The term v.h.f., therefore, is synonymous with very short wavelength.

Medium- and short-wave signals, about which we are probably more knowledgeable, do not always travel by a direct route from the transmitting aerial to the receiving aerial. These signals are reflected and bent by two electrically charged layers which surround the earth. The lower of these is sometimes called the Heaviside layer and the upper one the Appleton layer, named after their discoverers. They have no exact demarcation, but gradually occur and fade out with height above the earth, their height and electron density being influenced by the time of the year and the time of the day. The layers are composed of ionized particles and because of this the general term *ionosphere* is often used by engineers, and those of more recent years have given the depth of the ionosphere reference letters to identify the layers of ionization therein. The ionosphere acts to some radio waves very much like a mirror acts to light waves.

The signals radiated by the aerials of medium- and short-wave transmitters travel both skyward and also along the surface of the earth. The *ground wave* progressively diminishes in strength as the result of resistance set up by the earth's surface. Short-wave ground waves diminish in strength much more quickly with distance than do medium- and long-wave ground waves. This is the reason why good reception is rarely possible from a relatively local short-wave station. With ordinary local medium- and long-wave broadcasting, however, it is the ground wave which serves.

Now the waves which travel skyward (the sky waves) are reflected by the ionosphere and arrive back at earth at a distance from the transmitting aerial far in excess to that served by the ground waves. These layers, therefore, permit medium- and short-wave broadcasting over very great distances.

The diagram in Fig. 8.1 illustrates the paths that may be taken by waves from the transmitting aerial T. The receiving points R are shown placed at the positions of maximum signal due to the sky waves. It is also interesting to note that

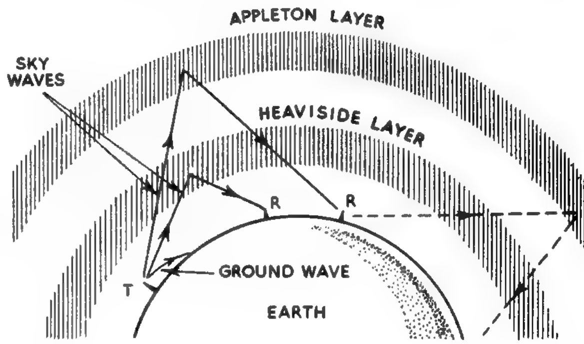


FIG. 8.1. *Paths that may be taken by the waves from the transmitting aerial T. The receiving points, R, are shown placed at positions of maximum signal due to the sky waves.*

the returning sky waves might well be reflected by the earth's surface and again sent skyward to continue the cycle (the paths shown in broken line). In this way extremely large distances are covered, and often the signal travels completely round the surface of the earth, in skips, several times before losing all its energy.

The influence that the ionized layers have on the signal depends not only on its frequency, but also on the time of year and the time of day. Short-wave signals tend to pass through the Heaviside layer and are usually reflected by the Appleton layer. Medium-wave signals, on the other hand, are reflected by the Heaviside layer, and since this is the layer nearest the earth the skip distance of medium-wave signals is correspondingly smaller than that of short-wave signals.

Distant medium-wave signals are much stronger after dark than during the hours of daylight owing to the considerably enhanced reflecting property of the Heaviside layer during the hours of darkness. Coupled with the progressive increase in number and power of medium-wave stations over the last few years, the factor of signal reflection is now making the reception of local medium-wave domestic broadcasting extremely difficult after dark. Whereas interference-free reception may well be possible during the hours of daylight, a little before dusk interference caused by frequency-sharing of more distant stations commences, and as the evening progresses the spurious signals often mar completely the

local station in many parts of the United Kingdom. Very-high-frequency f.m. broadcasting is, of course, the solution to the problem.

V.H.F. TRANSMISSION

Generally speaking, the service area where consistently good reception of any v.h.f. transmission can be obtained is almost confined to that area where the receiving aerials are in line-of-sight with the transmitting aerial. This is shown by the diagram in Fig. 8.2. Even if a line-of-sight path does not exist, the signal

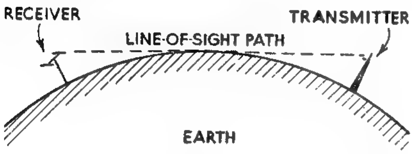


FIG. 8.2. *The service area where a line-of-sight path exists between transmitter and receiver.*

energy may still reach the receiving aerial. This is made possible by the earth's local atmosphere (known as the troposphere) refracting the ground (or "surface") waves so that for a while they follow the earth's curvature (Fig. 8.3). Any sky waves radiated at v.h.f. usually pass through both Heaviside and Appleton layers without reflection, and are thus lost in space.

For v.h.f. broadcasting, therefore, we have to rely on the rapidly diminishing ground waves. The area which falls beyond the line-of-sight distance of a trans-

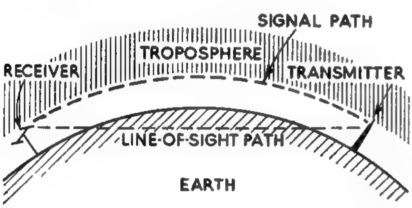


FIG. 8.3. *The fringe area where a line-of-sight path does not exist and where the signal path is bent by the influence of the troposphere.*

mitting aerial is referred to as the *fringe area*. The fringe area may extend approximately 30 miles beyond the *service area*, which is that area within the line-of-sight distance of a transmitting aerial.

LINE-OF-SIGHT FACTOR

As revealed by Fig. 8.2, the line-of-sight distance between transmitting and receiving aerials depends directly upon the height of the aerials above an arbitrary level on earth. It can be shown that the horizon distance in miles is approximately equal to 1·22 times the square root of the height of the observer in feet. Thus, the horizon distance from the top of a 100-ft. aerial is approximately 12·2 miles. This means that the maximum line-of-sight distance—where the line-of-sight path just skims the horizon—between the top of two 100-ft. aerials is twice the horizon distance, or 24·4 miles.

As another example, let us take a more typical case. Let it be supposed that the receiving aerial is situated 30 ft. above sea-level (an arbitrary level on earth). This gives an horizon distance of about $6\frac{1}{2}$ miles. If the transmitting aerial is, say, 1,300 ft. above sea level, the horizon distance here is about $43\frac{1}{2}$ miles. Now,

if the two horizon distances are added together, the maximum line-of-sight distance between the two aerials is 50 miles.

Under ideal conditions, assuming a hypothetical smooth spherical earth possessing little in the way of earthly features, such as irregularities of the ground, towns and large structures, such a distance could be considered as the line dividing the service area from the fringe area. In practice, though, the dividing line is in no way distinctive and it frequently falls at a distance less removed from the transmitting aerial than the line-of-sight distance. Apart from the above-mentioned factors, mineral content of the soil has an absorbing effect on the ground waves as they travel from transmitting aerial to receiving aerial.

TROPOSPHERIC EFFECT

Instead of the v.h.f. signals travelling parallel to the line directly between the transmitting and receiving aerials, they are bent slightly by the troposphere in such a way that they follow the curvature of the earth a little beyond the line-of-sight distance. They are not usually held close to the earth over great distances, however, since their energy is quickly dissipated.

As the troposphere is primarily the result of the variation of atmospheric pressure, temperature and moisture content with elevation above the surface of the earth, the bending effect is considerably influenced by weather conditions. In fringe areas, therefore, very inconsistent reception and serious fading of the v.h.f. signals might well be in evidence during a spell of unstable cyclonic conditions. During a spell of good conditions, however, strong steady signals are often received considerably beyond the range of the line-of-sight distance.

Freak v.h.f. reception is often possible over several hundred miles and signals from stations operating in shared channels are picked up by receivers at distances far in excess of the normal range, thereby causing serious local-channel interference. With f.m. broadcasting this is of little moment owing to the capture effect, but on television screens such interference can prove extremely disconcerting.

This effect was recently illustrated to the author when it was found possible, for several days, to obtain better reception from a Band III ITA television station situated 200 miles from the receiver than from the local Band III station situated 50 miles from the receiver. Both stations were using Channel 9, but it was impossible to view the local programme owing to interference given by the distant station. The aerial was re-orientated and the distant station was used until conditions became more settled.

As with television signals, very-high-frequency f.m. signals are reflected by passing aircraft and, in fringe areas especially, a "fluttering" disturbance often occurs on the reproduction during the time that the aircraft is in range of the aerial. The effect is caused by the aerial picking up two signals, the direct signal from the f.m. station and the reflected signal from the aircraft. Since the phase of the reflected signal is continually changing with flight of the aircraft, its energy contribution brought to the receiving aerial is alternately added to and subtracted from the energy of the direct signal.

The foregoing discussion clearly illustrates that the number of stations

required to provide a country-wide very-high-frequency f.m. service is considerably larger than the number of long- and medium-wave stations required to provide the same coverage. The BBC has progressively extended its coverage at v.h.f., concentrating first on the areas of unreliable long- and medium-wave reception, until approximately 98 per cent of the population of the United Kingdom has now been brought within easy reach of an f.m. station.

In such fringe areas as remain, securing the advantages of the f.m. system may demand the use of elaborate aerial arrays mounted as high as possible so as to increase the line-of-sight distance. Approximate coverage of the BBC f.m. stations is given in Fig. 8.4. It should be borne in mind that the screening effect of large buildings and hills is likely to cause local variations in reception.

Certain parts of the ionosphere can also reflect v.h.f. signals back to earth under severe sun-spot activity. The highest frequency that is reflected tends to vary on an eleven-year cycle, coincident with the sun-spot cycle, this usually being below about 25 MHz, meaning that higher frequencies are commonly passed into space through the ionosphere. When the sun-spot activity is at its height, therefore, and at some other times, the ionosphere can become so "electrically" dense as to reflect signals well above 30 MHz, well into the v.h.f. spectrum, thereby preventing them from entering space.

Since the ionosphere is at a greater altitude above earth than the troposphere, ionospheric propagation results in the return of signals over great distances.

F.M. AERIAL SYSTEMS

Unlike ordinary a.m. receivers, very-high-frequency f.m. receivers invariably demand the use of aerials of special design. The aerials are self-tuned to the frequency corresponding approximately to the mean of Band II by virtue of their electrical length. We can, in fact, consider aerials of this kind as tuned circuits, as distinct from the odd length of untuned wire which is nearly always found serving a.m.-only receivers. When a tuned aerial receives a signal falling within its response, the resulting voltage and current are not evenly distributed along its length.

For example, with a rod whose length is half of the wave-length of the required signal, the voltage maxima occurs at the ends, while at the centre the voltage is at a minimum. Conversely, the current is at a maximum at the centre and minimum at the ends. As would be expected, the voltage and current distribution follows the usual variations associated with an alternating quantity, as shown on the diagram at Fig. 8.5. It will be recognized that there exists a phase displacement of 90 deg. between the voltage and current. These voltage and current components due to the signal comprise *standing waves*.

Since an aerial is concerned with both voltage and current, it follows that it must possess impedance. With an aerial a half-wavelength long the impedance is at its lowest at the centre, which would follow since this is also the point of maximum current. At the ends, at maximum voltage, the impedance is also at maximum. Although it may seem that the impedance is zero at the centre, this is not so in practice; here it is in the region of 70 ohms and rises to some thousands of ohms at the ends.

It is general practice to divide the half-wave rod or aerial into two parts with

FIG. 8.4. *Service areas of the BBC f.m. system. The frequency and power of each of the transmitting stations are shown in Table 1.1 on page 16.*

an air-gap of about 1 in. between them. A special insulator is often employed to perform this function while also endowing the aerial with mechanical stability along its length.

Apart from the insulator being of use to secure a mast to the aerial—or *dipole*, as it is now known—without having an effect on the tuned element, it also serves to permit connexion of the dipole at its low-impedance point to the cable or

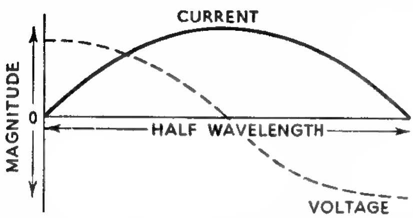


FIG. 8.5. *The voltage and current distribution along a half-wave dipole aerial. The centre point, of maximum current corresponds to the point of minimum impedance. At the ends the current is minimum and the impedance maximum.*

feeder used for carrying the signal picked up by the dipole to the receiver. Maximum transfer of signal occurs only when the impedance of the feeder equals the impedance of the connecting point on the dipole and, similarly, when the impedance of the feeder equals the input impedance of the receiver.

From the aerial point of view, a feeder possessing an impedance of between 70 and 80 ohms provides an optimum match to the centre of the dipole, while at the receiver end a transformer is used which steps up the impedance of the feeder so that a correct load is reflected to the input circuit of the r.f. valve.

The sketch, Fig. 8.6, illustrates how a simple f.m. dipole may be constructed from two pieces of wood joined in the form of a T, two lengths of metal rod or

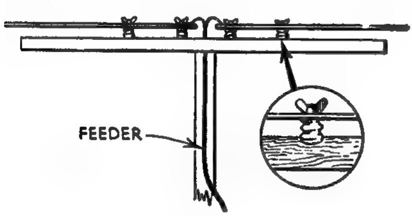


FIG. 8.6. *Simple f.m. aerial constructed of two pieces of wood joined in the form of a T, two lengths of metal tube each 2 ft. 6½ in. long, and four insulators.*

tube, each 2 ft. 6½ in. long, and four stand-off insulators. The parts should be assembled as shown, and the rods arranged so that the spacing at the centre is a little under one inch. The feeder should be connected as close as possible to the ends of the rods, preferably by soldering or, alternatively, the two halves of the feeder may be clamped beneath the wing-nuts securing the dipole to the insulators.

Since the f.m. signal is horizontally polarized the aerial must also be mounted horizontally. Owing to this the aerial exhibits directional properties, and maximum signal is induced when the dipole is facing broadside to the transmitting aerial. Thus, a simple dipole mounted horizontally is said to have a "figure-of-eight" polar diagram; that is, the signal pick-up becomes less and theoretically falls to zero as the aerial is rotated over 90 deg.—in relation to the transmitting aerial—from the broadside position to the end-on position. If the rotation is continued, the signal pick-up gradually increases until, again, the dipole is facing broadside to the transmitting aerial.

It is interesting to note that a simple dipole mounted vertically, in order to respond to vertically polarized television signals, for example, will pick up signals equally well from any point of the compass, and is for this reason termed *omnidirectional.*

FEEDERS

Two kinds of feeder are extensively used with f.m. and television receivers. Balanced feeder, having a characteristic impedance of 70 to 80 ohms, consists of two parallel wires spaced about 1 mm apart, and moulded together in a low-loss insulating material (dielectric) produced of telcothene or polythene. The whole is sometimes enclosed in a copper mesh and is then known as screened balanced feeder.

An unbalanced type of feeder, generally known as coaxial or concentric feeder, represents the other popular type of transmission cable at v.h.f. This consists of a central conductor, either stranded or solid, moulded in polythene and surrounded by a copper mesh which also serves as the second conductor. The whole is enclosed in a weatherproof covering of polyvinyl chloride or similar material.

Both kinds of feeder are made in various grades and characteristic impedances. The most popular values for f.m. and television use range within 70 to 80 ohms. Since feeders will be subject to losses the signal at the set end of the feeder is always less than the signal at the aerial. Moreover, the higher the frequency of the signal the greater the losses in the cable. For this reason the feeder run should always be kept as short as possible at v.h.f. and on no account should surplus feeder be coiled up behind the receiver.

The semi-air-spaced type of coaxial feeder is very useful in fringe areas of low signal strength since it has a low loss at v.h.f. This feeder is more expensive and larger in diameter than the standard grade, but it is well worth using in cases where an extra microvolt or so of signal at the receiver may make all the difference between fair reception and good reception.

One should always be careful to select the type of feeder for which the receiver was designed. European and American manufacturers favour 240–300-ohm balanced aerial circuits, while British makers prefer 75-ohm unbalanced circuits to cater for "standard" TV coax. The use of coaxial feeder on a receiver designed for balanced feeder is often successful within the service area and at locations of high signal strength, but such practice in fringe districts might well result in inferior reception coupled with an unnecessarily high level of interference. The same reasoning applies with respect to the use of balanced feeder on receivers designed for coaxial feeder.

Transformers are available, however, for insertion between, say, a receiver designed for balanced feeder and the coaxial feeder from the aerial. They can be employed in the reverse sense so that balanced feeder from the aerial can be used with a receiver designed for coaxial feeder. Although it is desirable to use the correct type of feeder (and aerial with a matching impedance, of course), a balun transformer can be used to match 75-ohm coaxial into a 300-ohm balanced input (see Chapter 4); but care must be exercised to ensure that such a transformer does not introduce excessive loss, especially where the receiver is to be

used for stereo in a poor signal field area. My companion volume, *The Practical Aerial Handbook* gives details of a suitable low-loss design. It might also be possible to convert a 300-ohm balanced input to 75 ohms unbalanced. This was explained in Chapter 4.

It is sometimes believed that the use of unscreened balanced feeder, in favour of coaxial feeder, aggravates the interference problem. This is not the case because, although unscreened feeder picks up more interference it is true, the interference induced in one conductor is equally as strong and of the same phase as the interference induced in the other conductor.

At the receiver, therefore, the interference signals are applied across opposite ends of the primary winding of the aerial-coupling transformer; since they are of equal strength and of the same phase they cancel out and little or no interference is induced across the secondary of the transformer. The success of this artifice for alleviating interference depends upon the accuracy of balance of the aerial-input circuit. Owing to the effects of spurious capacitances it is rarely feasible to secure one-hundred-per-cent balance on commercial receivers at v.h.f. It is, of course, pointless to endeavour to achieve the effect by using balanced feeder on receivers designed for coaxial feeder, since the aerial input circuits here are purposely unbalanced to match that characteristic of the feeder.

Balanced feeders are, therefore, recommended for use as down-leads in areas of high interference, even if this practice demands the employment of a balun transformer; screened balanced feeders are even more desirable under such conditions, but every effort must be made to mount the aerial above the field of interference. When screened balanced feeders are used, the copper-mesh screening should be connected to an *efficient* earth point at the earth socket of the receiver.

DIPOLE PLUS REFLECTOR

By mounting another slightly longer rod at half, quarter or one eighth of the signal wavelength behind the dipole, the array is made even more directional. This arrangement of dipole plus reflector is often referred to as an H-type aerial, and is extensively used, mounted vertically, for the reception of television signals.

Because the reflector, as well as the dipole, is excited by the signal, the reflector signal is re-radiated and is picked up by the dipole in phase with the signal picked up by the dipole itself. In this way the reflector signal and the dipole signal are added together. This results in almost total cut-off of signals arriving at the reflector side of the array, while reception of signals from the opposite direction is increased almost twofold over a simple dipole. Put a little more technically—the reflector increases the pick-up efficiency in the forward direction by about 5 decibels, as compared with a simple dipole, and reduces it in the rear, giving a front-to-back ratio of some 9 decibels.

The effect can be made even more marked by the addition of another rod of slightly less length than the dipole, placed 0·1 wavelength in front of the dipole. This extra element is known as a director and serves to make the array extremely directional, giving a front-to-back ratio of some 18 decibels while also increasing the forward pick-up efficiency by about 8 decibels over a simple dipole.

Figs. 8.7 and 8.8 respectively, show two- and three-element f.m. aerials. The two-element type is popular in districts towards the edge of the service area. The three-element array on the other hand is suitable for use in fringe districts and in those small "pockets" in service areas where reception may be difficult because of adverse local conditions.

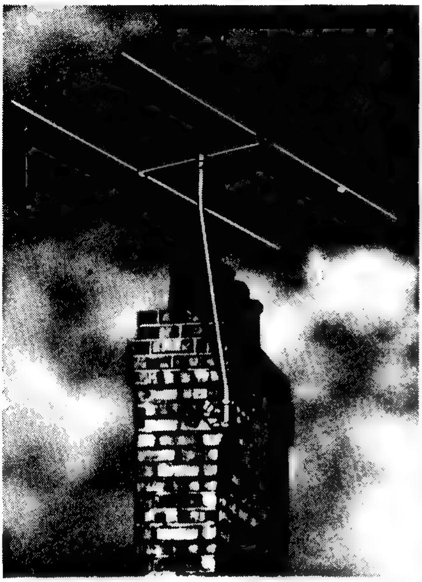


FIG. 8.7. *Two-element f.m. aerial.*

FIG. 8.8. *Three-element f.m. aerial.*

In some service areas a Band I television aerial will often provide satisfactory Band II signals, but clearly, owing to the difference between the lengths of the elements and polarization of the signals, this is far from the most efficient way of securing f.m. signals.

To preclude the necessity of two or, in some cases, three outside aerial systems, horizontal Band II stubs are available whose purpose is to adapt an existing television aerial so that it becomes properly responsive to Band II signals. Adaptors of this nature are illustrated in Fig. 8.9 in conjunction with a simple Band I dipole.

COMBINING AERIAL OUTPUTS

With composite aerials the signals in all bands are carried by a single feeder to the receiver. No problems are presented when such an aerial is used with a television v.h.f./f.m. receiver, for invariably this kind of receiver features a single aerial socket which is common to all bands, and the receiver switching selects the required signal.

Where conditions make it necessary to use separate aerials, however, a problem is presented. The feeders can, of course, be changed over when the band-switch on the set is changed, but this is inconvenient and might lead to premature

failure of the receiver's aerial socket. It is not a good policy to join the feeders of three aerials (Band I, Band II and Band III) working at different frequencies, because interaction between the signals will occur and detract from the strength of the signals in individual bands, apart from the factor of the resulting mismatch at the receiver input.

These undesirable effects can be avoided by combining the signals in a triple filter network having high-pass, band-pass and low-pass sections. If the Band III television signals are taken through the high-pass section, the Band II f.m.

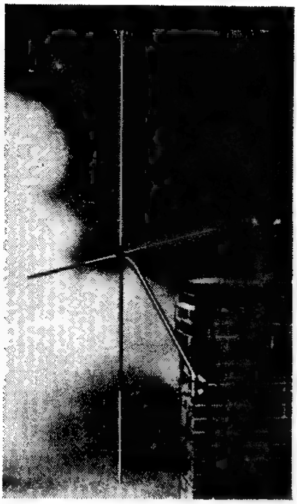


FIG. 8.9. *F.M. adaptor kit used in conjunction with a Band I dipole.*

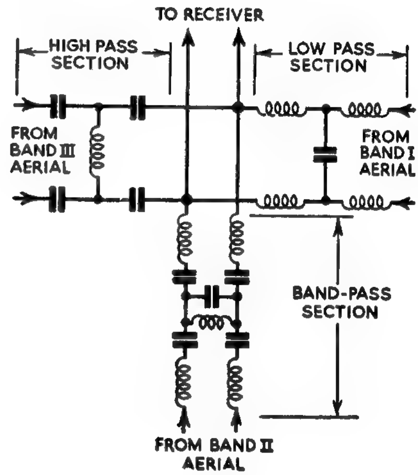


FIG. 8.10. *Circuit of a high-pass/band-pass/low-pass filter used to combine the signals in Bands I, II and III to a single feeder.*

signals through the band-pass section and the Band I television signals through the low-pass section, to the receiver's common aerial socket, very little attenuation will occur on the signals passing from the aerials into the filters, but considerable attenuation will be offered to the frequencies of the signals involved across the outputs of the filters. A circuit of a high-pass/band-pass/low-pass filter of this kind is shown in Fig. 8.10. Commercially, these filters are often of printed-circuit design and are known as Triplexers.

It is possible to use them in the reverse sense, so that a composite aerial having a single feed can be connected to a television receiver having separate input sockets for Band I and Band III and to an independent f.m. receiver or tuner (Fig. 8.11). Where there is a considerable distance between the aerials and the receiver(s), instead of using three runs of feeder, it is sometimes more economical to use two filters, one which has been made waterproof situated near the aerials, so that their outputs can be combined to a common feeder, and another near the receiver(s), so that the signals can be split for application to separate input sockets (Fig. 8.12).

The slot aerial may be used as an alternative to the dipole and reflector or simple dipole, to which it has approximately similar performance. It functions

differently, however, in that it produces and responds to a horizontally polarized field when it is vertically positioned, in converse to the dipole which produces and responds to a vertically polarized field when it is vertically positioned. In other words, the electric and magnetic fields are interchanged when changing

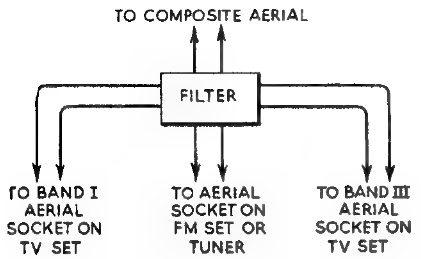


FIG. 8.11 (left). *Connexions for a single filter unit, often known as a Triplexer.*

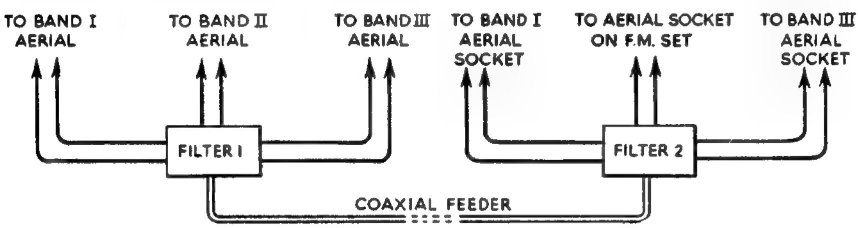


FIG. 8.12 (below) *shows the use of two separate filters as explained in the text.*

from a dipole to a slot. It will be recalled from Chapter 1 that the direction of polarization of an electro-magnetic field is that in which the electric field lies. In order to respond to the f.m. signals, therefore, a slot aerial will need to be positioned so that the long side of the slot is in the vertical plane.

The sketch, Fig. 8.13, illustrates the construction of this simple kind of aerial. The slot is about a half wavelength long and is made in a sheet of conducting

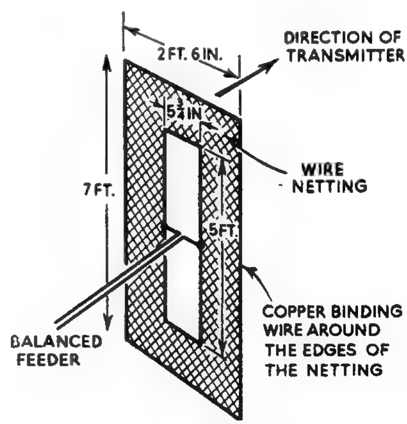


FIG. 8.13. *Construction of a simple slot dipole.*

material, such as wire netting. As with the horizontal dipole, the vertical slot also has a radiation pattern which is in the form of a figure eight. This means that it is sensitive to signals arriving in either direction at right-angles to the slot. Owing to its nature of construction, it is not very suitable for mounting out-of-doors, but is often of value for mounting in lofts.

Although the characteristic impedance between the centre points of each of the vertical sides is in the region of 600 ohms, the degree of mismatch by the use of 75-ohm balanced feeder does not appear to impair the performance of the aerial to any great extent.

Fig. 8.14 shows the best method of connecting coaxial feeder to the slot. A length of 18 s.w.g. copper wire is held by insulated spacers in the centre of one

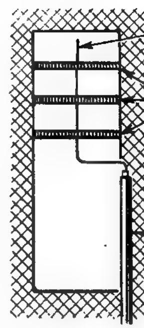


FIG. 8.14. *The best method of connecting coaxial feeder to the slot.*

half of the slot. This loading element is connected to the centre conductor of the coaxial feeder, which is secured electrically by its outer braiding along the edge of the slot, as shown.

An array of slot aerials is used by the BBC for radiating very-high-frequency f.m. signals. Fig. 8.15 shows the v.h.f. aerial at Wrotham. The aerial consists of

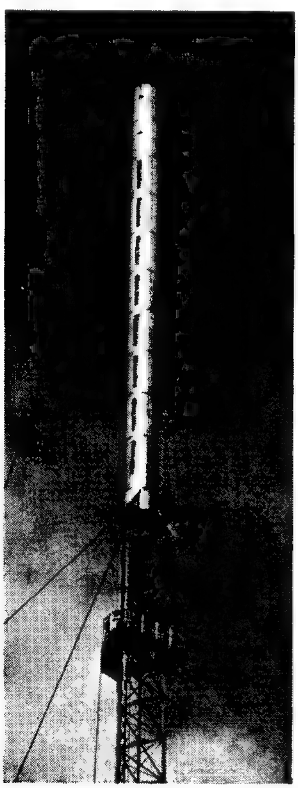


FIG. 8.15. *The v.h.f. aerial at Wrotham, Kent.*

32 slots in the wall of a cylinder 110 ft. long and 6½ ft. in diameter. The slots are arranged in 8 tiers with four slots in each tier.

AERIALS FOR STEREO

Owing to the wider frequency spectrum of stereo—extending to 53 kHz as distinct from the 15 kHz of mono—and the action of the decoding, where the signal is effectively sampled over the two a.m. envelopes (see Chapter 7), the noise yield of a receiver working stereo can be significantly above that of the same set working mono, depending on the applied aerial signal strength and design of the system. It is often said that the signal/noise ratio is worsened on stereo by as much as 20 dB when the signal is just about strong enough to give the maximum noise performance of the receiver on mono. In practice, the best of tuners and decoders exhibits a lesser signal/noise ratio impairment. I have, indeed, checked several such systems in which the mono signal/noise ratio is about —40 to —35 dB when the aerial signal is just strong enough to take the tuner towards full amplitude limiting (and in some tests this represents a signal of a little more than 20 μV!) and on switching to stereo (with a stereo transmission, of course) the signal/noise ratio has dropped only to about —32 to —30 dB. In other words, this implies that the tuner/decoder set-up is sufficiently sensitive to give fair stereo reception on an aerial signal as low as 2 μV to 5 μV. For this outstanding performance a well designed decoder has to be partnered with a tuner of excellent noise features, using one or two f.e.t.s in the front-end (see Chapter 4). On systems of lesser exactitude the stereo noise at such a low input can render comfortable listening totally impossible, clearing when the mono/stereo switch is flicked to mono.

Having said all this, however, the importance of the aerial system cannot be overlooked, for even though the set noise may be tolerable impulsive interference can rise to uncomfortable levels on stereo due partly to the a.m. subchannel carrying the stereo information. It must also be remembered that the amplitude of the pilot tone is considerably below that of the other multiplex signal components, and under certain conditions this can be lost below noise and with it the stereo effect. This is less likely to happen, though, with decoders featuring pilot tone limiters. It has been suggested (by A. H. Uden in "FM Diary" of *Hi-Fi News*, Volume 11, No. 7, December, 1966, page 691) that the suitability of the aerial for stereo can be tested by inserting 12 dB of attenuation in the aerial feeder to the set when the set is running mono. If satisfactory signal/noise and signal/interference ratios are still maintained with this degree of attenuation, it can *generally* be assumed that the receiver will function in a like manner on stereo. It is noteworthy that some of the very latest stereo tuners and tuner-amplifiers can be operated virtually down to noise performance roll-off on mono *and* stereo alike with acceptable reproduction.

Multipath interference can also impair the stereo performance since the phasing of the pilot tone in the reflected signals might differ from that of the tone in the direct signal. The regenerated subcarrier can then get into trouble with synchronization, hunting between one signal and the other. Differential fading in adverse reception areas can also cause the subchannel signal components to appear out of correct amplitude ratio or, indeed, slightly out of phase due to

phase distortion effects over the signal path and during the abstraction of the signal energy by the aerial. Some experimenters have discovered that by "beaming-off" the main signal to secure improved discrimination against unwanted adjacent channel signals or reflections of the wanted signal an "edginess" is given to the stereo reproduction. This could well be caused by phasing disturbances as just mentioned.

When stereo signals are present in adjacent channels (200 kHz apart) and one signal is many times stronger than the other, reception of the weaker signal is sometimes marred by warbling effects or "birdies", as described in Chapter 7. This can happen, too, on a "local" stereo signal when an adjacent channel also carries a stereo signal some magnitudes below the "local" one. Filtering the interbeat components of the two subchannels can sometimes effect a cure, as mentioned in Chapter 7.

Finally, a word about long-distance reception, usually called DX reception. It is often possible with a very sensitive tuner to obtain fair reception of mono

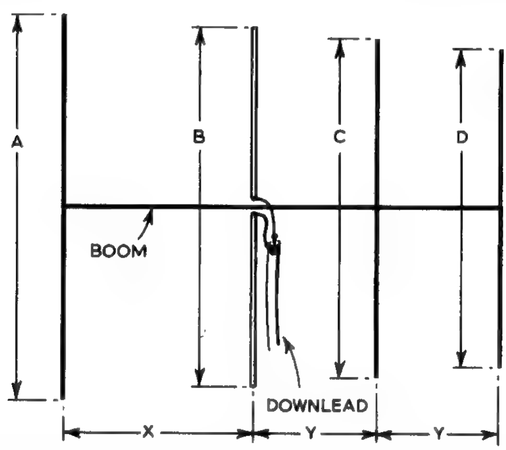


FIG. 8.16. *Design for a v.h.f./f.m. four-element aerial, based on information supplied by the BBC. Dimensions are: A 5 ft. 4 in., B 5 ft. 0 in., C 4ft. 9 in., D 4 ft. 6 in., X 2 ft. 8 in. and Y 1 ft. 7 in. The feeder connecting points on the dipole should be about 1 in. apart and insulated from the boom. Note that the dimensions given for the elements (not the spacings) are values of length* plus *diameter, so if the elements are made of, say, ½ in. rod, the correct lengths are ½ in. less than detailed above. Based on information supplied by the BBC.*

and stereo over paths of a hundred miles or more, depending on the tropospheric conditions and the time of the year. Early spring is a good time for this sort of reception, and at the time of writing I am getting stereo from stations up to 300 miles distant. High-gain f.m. aerials are available for reception in distant and difficult locations, a good example being the range of stereo f.m. aerials (up to six elements) by J-Beam Aerials Ltd. Some of these are tuned for optimum reception of the Wrotham Radio 3 transmission. Aerial amplifiers and set-side f.m. boosters can also be useful where the aerial signal is particularly weak so as to amplify it sufficiently to push the set into amplitude limiting. Details of boosters like this and high-gain aerials are given in my companion volume, *The Practical Aerial Handbook*, 2nd Edition. However, for experimenters wishing

FIG. 8.17. *Wenvoe aerial details (BBC Photo).*

FIG. 8.18. *More feed details of the Wenvoe aerial (BBC Photo).*

to try their hand at making a four-element f.m. aerial design, particulars are given in Fig. 8.16. Also as a matter of interest Figs. 8.17 and 8.18 show feeder details of the BBC TV and v.h.f. transmitting station at Wenvoe, South Wales. The first gives a view at the bottom of the v.h.f. cyclinder (see Fig. 8.15, for example), revealing the splitter and horizontal feeders, while the second shows the feed-plate at the aerial end of the distribution feeders—somewhat more massive than found in the domestic installation, to say the least!

Let us end this chapter with a word from the BBC about long distance reception. "We have two thoughts about this, one is that because of the 'cussedness' of Nature good conditions are not usually present when the programme you greatly require is being transmitted which, to say the least, is highly frustrating; and secondly, a broadcast engineer usually takes a very jaundiced view of long distance reception since this usually means additional protection or increased curtailment of the number of transmitters that can be planned to operate in any particular broadcast band. Nevertheless, having said that we would, of course, agree that a fraction of a loaf is usually better than none". In other words, one just cannot rely on reception of v.h.f. signals (or u.h.f. signals for that matter) outside the planned-for service areas.

CHAPTER 9

F.M. Receiver Alignment

THERE are two methods of aligning f.m. receivers and tuners. By far the most accurate method makes use of a wobbulator and oscilloscope, and possibly an accurately calibrated signal generator, but in many professional and amateur workshops, although a signal generator may be available, a wobbulator and oscilloscope are not always at hand. In such cases the somewhat less accurate method has to serve.

This method requires the employment of a signal generator covering Band II and the intermediate-frequency of the receiver undergoing adjustment, and in addition a sensitive high-resistance voltmeter. This method will be described first.

SIGNAL GENERATOR AND VOLTMETER METHOD

As is common with a.m. receivers, the lining-up process starts at the i.f. stages. It is first necessary to concentrate on the pre-detector i.f. stages; that is, leaving the *secondary* of the discriminator transformer until last.

It is, in fact, often necessary to detune the secondary of the discriminator transformer by unscrewing its core until it is well away from the winding. This unbalances the discriminator circuit and, for certain receivers, aids in the production of an indication of resonance in relation to adjustments of the i.f. transformers.

For initial adjustment of the i.f. stages, the voltmeter is connected across the R/C load in the case of a ratio detector circuit, as shown in Fig. 9.1. One side of this load connects to chassis, so it is necessary simply to connect the voltmeter between chassis and the other side of the load, with the positive side of the voltmeter connected to chassis.

Where a balanced ratio detector circuit is featured, the voltmeter can still be connected to the same point, as shown in Fig. 9.2, but here the voltage across one half (R2) of the resistive load (R1 + R2) will be applied across the voltmeter. This, however, is of little consequence and will not affect the alignment in any way.

The "live" lead of the signal generator is connected to the control grid of the final i.f. amplifier valve, i.e. the valve immediately before the ratio detector, and the screen to chassis (for transistor receivers see page 176). In certain

receivers, particularly those whose design demands the connexion of the chassis to one side of the mains supply, the signal generator connexion should be made through 0·001 μF capacitors, so as to isolate the generator output circuits from the receiver chassis. In addition, on the type of receiver mentioned, it is most

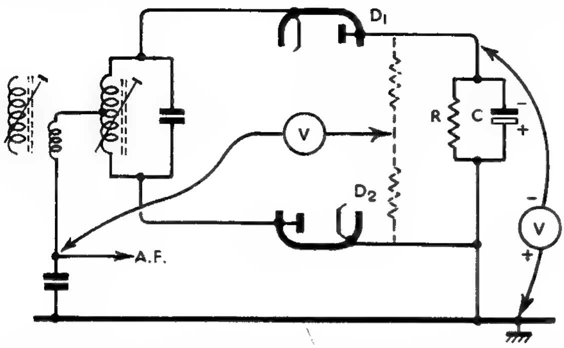


FIG. 9.1. *Showing the positions of voltmeters in an unbalanced ratio-detector circuit to act as tuning indicators. The resistors shown in broken line are temporarily included to facilitate the balancing adjustment.*

desirable to ensure that the mains supply is applied to the receiver so that the chassis is connected to the neutral side of the supply. This precaution should be exercised during any servicing operation, solely as a means of ensuring that the receiver chassis never becomes "live" with respect to earth. A neon tester held on the chassis when the set is energized is the best way of establishing that a chassis is "dead"—the neon will glow, of course, if the chassis is "live", and in this case the mains plug should be reversed.

The signal generator should be allowed to warm up properly and thus become frequency stable; this usually takes ten to fifteen minutes. With the modulation switched off, the generator should be adjusted to the receiver's intermediate frequency, and the core in the *primary* of the discriminator transformer should be adjusted for maximum reading on the voltmeter.

Since only one i.f. stage is at present utilized, it may be found necessary to inject quite a large signal in order to secure a reasonable deflection on the meter, but the reading is usually about 20 to 30 volts on a meter having a sensitivity of 10,000 ohms-per-volt. A meter of lower sensitivity can be used for this application, and quite successful results have been achieved by the author using a meter of 1,000 ohms-per-volt. The higher the resistance of the meter the less will be the loading on the detector.

The signal should now be transferred to the control grid of the next i.f. valve in line, going back towards the aerial, and the cores in the associated i.f. transformer (not the one in the grid circuit of the valve to which the signal is applied) should be adjusted for maximum reading on the voltmeter, adjusting first the primary and then the secondary cores of the transformer.

This process should be continued until the first i.f. transformer in line has received attention, always adjusting first the primary and then the secondary. In order to adjust the first i.f. transformer—that is, the one with the primary in the anode circuit of the mixer valve—the signal will have to be applied to the control grid of the mixer valve. Usually, it is recommended that the signal is very loosely coupled to this circuit and on commercial receivers, including those catering for both a.m. and f.m., a point is usually available on the tuner unit

section to facilitate connexion of the signal without undue disturbance to the associated circuits.

If such feature is not embodied, however, a length of copper wire wound round the envelope of the mixer valve and connected to the live lead of the generator provides sufficient coupling of the signal. On receivers which use a screening can round the mixer or frequency-changer valve the signal can be connected to this can after it has been slid up the valve so as to disconnect it from the chassis.

The gain of the i.f. section of the receiver at this point will be very high, so the signal input should be reduced accordingly, and this also applies as the generator is progressively transferred from grid to grid along the i.f. channel.

It now remains to tune the secondary of the discriminator transformer, and this operation is probably the most important of them all, for unless an accurate balance is achieved in the two halves of the detector circuit, not only will the reproduction be far below the standard expected of f.m. equipment, but any amplitude disturbances present on the carrier will not be eliminated, and excessive interference is likely to be present on the a.f. signal.

Reverting to the balanced ratio detector circuit in Fig. 9.2. The circuit is in balance when the diode currents are equal; this was discussed in Chapter 3.

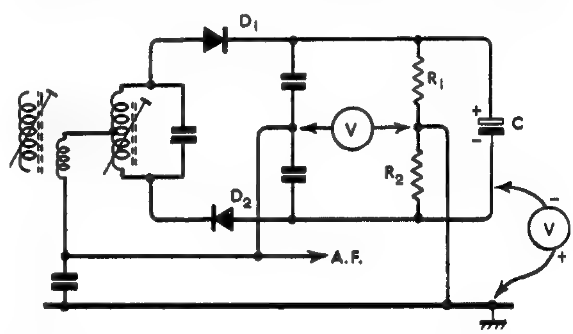


FIG. 9.2. *How the voltmeters are connected in a balanced ratio-detector circuit. The meter connected at the junction of R1 and R2 and the a.f. take-off point should have a high resistance.*

When this state exists, the voltage difference between the junction of the load resistors R1 and R2 and the centre-tap of the secondary of the discriminator transformer will be zero. Thus, there is a simple means of checking for balance, and making the necessary adjustment, by connecting a meter between the points mentioned, as shown in Fig. 9.2. The meter cannot be connected to the tap of the secondary, but from the d.c. point of view connexion to the a.f. feed-off point amounts to the same thing. The other side of the meter can, if more convenient, be connected to chassis, since the junction of R1, R2 is returned to chassis anyway.

With the meter so connected and a signal at i.f. applied to the i.f. channel, as previously described, the secondary of the discriminator transformer should be carefully adjusted for *zero reading* on the meter. This will occur mid-way between a negative-going and a positive-going deflection.

With the unbalanced circuit (Fig. 9.1) there is not a convenient centre-tap on the load resistor, so purely for the purpose of performing the balancing adjustment a *matched* pair of resistors is introduced into the circuit temporarily, as

shown in broken line on the diagram. The meter is then connected between the junction of the resistors and the a.f. take-off point, as in the previous case. The resistors should each have a value of about 47,000 ohms, and it is most important for this application to employ a meter of very high resistance. A valve-type voltmeter is desirable, but an ordinary meter of about 5,000 to 10,000 ohms-per-volt sensitivity is generally suitable. If a voltmeter for this purpose is not available either a 50-microamp. meter connected in series with a 100,000-ohm resistor or a 100-microamp. meter connected in series with a 47,000-ohm resistor can be used instead.

The foregoing description of aligning the i.f. circuits is applicable to nearly all receivers featuring a ratio detector and, in most sets, the procedure of merely peaking the i.f. transformers for maximum output provides for a reasonable discriminator characteristic, in spite of the simple nature of the adjustments. Naturally, the exact method of aligning depends on the instructions given for that particular model in the maker's service manual and, where possible, this should always be procured and the instructions faithfully followed.

In some receivers the second i.f. transformer is "double-humped" (tightly coupled) and the primary and secondary are designed to respond to different frequencies. Information of this nature can be acquired only from the service manual so it is not possible to generalize here, but where such a transformer is used, a damping resistor should temporarily be connected across the winding which is not undergoing adjustment.

There are one or two other ways of aligning the i.f. stages and discriminator transformer, even without a wobbulator and oscilloscope, but they are probably not as effective as the method discussed.

By detuning the secondary of the discriminator transformer the detector will respond to amplitude modulation, so by using an a.m. signal generator and output meter connected across the loudspeaker, the i.f. stages can be aligned, according to ordinary a.m. practice, by adjusting for maximum output on the meter. The secondary winding can be finally adjusted by tuning for *minimum* output on the meter, thus indicating maximum a.m. rejection and balance of the ratio detector.

In addition, an f.m. generator can be used to peak the i.f. transformers, again, aided by an output meter across the loudspeaker, but this method is not often recommended as some other method has to be devised to balance the discriminator, in any case.

One or two words of warning. Where the receiver or tuner features a limiter, the input signal should always be kept low enough to avoid limiting action consistent with workable indication on the meter. An excessive signal will promote heavy limiting action and might well mask the effect of adjustments of the i.f. transformers.

Bear in mind that an i.f. bandwidth approaching 200 kHz is necessary to allow unrestricted passage of the sideband components of the f.m. signal. In some cases, therefore, particularly if the reproduction is lacking in top, after peaking the i.f. transformers, it may be found desirable to detune them slightly to ensure the desired overall response. If this is found necessary, one should always finally ensure that the detector still remains properly in balance.

F.M. RECEIVER ALIGNMENT

A rough idea of the performance of the receiver and the symmetry of the detector can be gleaned by observing the effects of tuning over a station. Detuning either side of a station will result in severe distortion, owing to rectification of the f.m. signal occurring on a non-linear portion of the discriminator curve. If such distortion, of equal magnitude, occurs the same distance away from the station on either side of the correct tuning point, then it can be assumed that the response is reasonably symmetrical. Distortion-free reproduction should, of course, be obtained when the tuning is accurately adjusted. A low-level distortion-free output will be obtained on either side of the correct tuning point, before the areas of heavy distortion, as the result of the signal being rectified fairly linearly by the side responses of the S-shaped discriminator characteristic.

ALIGNING THE FOSTER-SEELEY CIRCUIT

Without the assistance of an oscilloscope and wobbulator, it has been found in practice that alignment of the i.f. and discriminator sections of a receiver using a Foster-Seeley circuit can best be effected successfully simply by the use of an a.m. signal generator and a.f. output meter.

After first detuning the discriminator by unscrewing the core adjustment of the secondary winding of the discriminator transformer, the i.f. transformers are all adjusted for maximum output on the meter connected across the loudspeaker. The a.m. signal from the signal generator, at the accurate intermediate frequency, is applied to the various stages, as described in the former case. Again, care should be taken to ensure that the limiter is not brought into operation, by keeping the generator signal at the lowest possible level. It should be remembered that an amplitude limiter stage is almost certainly to be employed in receivers where a Foster-Seeley discriminator is used in favour of the ratio detector.

After adjustment of the i.f. transformers, without altering the generator frequency, the core adjustment of the secondary winding of the discriminator transformer is screwed in and very critically set for *minimum* reading on the output meter, if necessary increasing the signal input to give an adequate reading.

Certain receivers may require a damping resistor connected across the secondary of the second i.f. transformer (a value in the region of 39,000 ohms is typical) while adjustments are made to the i.f. transformers, but this should be removed before adjusting the secondary of the discriminator transformer.

ALIGNING THE OSCILLATOR AND R.F. SECTIONS

Alignment of the oscillator and r.f. sections of an f.m. receiver should not present undue difficulty, and the process is aided considerably by the use of an f.m. signal generator which is tunable over Band II.

Before any adjustments are attempted, the receiver or adaptor should be allowed to warm up for at least ten minutes to permit stable operation of the local oscillator. The tuning pointer or cursor should also be adjusted so that it traverses over the calibrated section of the tuning scale between the maximum and minimum settings of the tuning gang or ganged inductors.

On most circuits it is required only to align at one spot frequency at about

93 MHz; a calibration mark is sometimes provided on the scale for this purpose. The f.m. signal (modulated to 30 per cent) is applied to the receiver aerial socket and an a.f. output meter is connected across the loudspeaker. The oscillator trimmer is then adjusted for maximum output on the meter and the generator signal reduced until the smallest workable indication on the meter is obtained. The aerial and r.f. trimmers are then adjusted for maximum output on the meter, if necessary reducing the signal input in order to maintain the smallest possible deflection on the meter.

Some sets require alignment at both the high-frequency and low-frequency ends of the band. On such sets it is often necessary to align at 88 MHz and at 95 MHz, adjusting first the tuning slugs in the appropriate coils and then the parallel capacitor trimmers, if fitted, as a means of achieving optimum tracking, consistent with even sensitivity, over the whole of the band.

Where an f.m. signal generator is not available, the above procedure may be followed using the local BBC transmissions. Where it is necessary to align at only one spot frequency, the Radio 3 transmission should be selected, while where alignment at two points is required, the Radio 2 and Radio 4 programmes should be chosen, since these represent the lowest and highest frequencies available, while Radio 3 has a frequency between the two.

In areas of high signal strength it may be found necessary to attenuate the aerial signal when making adjustments to the aerial and r.f. circuits to prevent operation of the limiter masking slight increases in receiver sensitivity as the circuits are brought into tune.

Before a discussion of the visual method of alignment can be continued it will be necessary to learn a little about the oscilloscope ('scope for short) and the more modern version of the wobbulator.

THE OSCILLOSCOPE

The basis of any 'scope is an electrostatically focused and deflected cathode-ray tube. Most readers have a fair understanding of this apparatus and are familiar with its physical make-up, if not in relation to servicing equipment, at least from the aspect of television.

The high-speed beam of electrons produced by an assembly of electrodes, situated in the neck of the tube and known as the electron gun, impinges on the fluorescent screen. This screen possesses the curious property of changing the stored (kinetic) energy of the electrons, derived in virtue of their mass and velocity, into light radiation. Other electrodes bring the electron beam to a sharp point of focus at the fluorescent screen, while situated between the gun and the screen are two pairs of metal plates, known as deflecting plates. These make it possible to deflect the electron beam both in horizontal and vertical directions across the screen. The extent of deflection depends on the degree of voltage applied across the plates.

The plates which are concerned with deflecting the beam horizontally are known as the X plates, while the plates which deflect the beam in a vertical sense are known as the Y plates.

For a large number of applications of the 'scope the electron beam is arranged linearly to traverse the screen of the c.r.t. in a horizontal direction (from left to

right), while the voltage waveform under examination is arranged to deflect the beam vertically.

A timebase inside the 'scope gives the function of horizontal deflection by subjecting the X plates to a sawtooth voltage, which rises linearly with time and moves the spot of light on the screen, produced by the electron beam, from left to right. At the finish of the rising side of the sawtooth voltage, the spot of light is at the right-hand side of the screen, but it swiftly returns to the left-hand side to recommence the horizontal trace as the voltage of the X plates follows the rapid fall of the trailing edge of the sawtooth voltage waveform.

The timebase has facilities for varying the repetition frequency of the sawtooth waveform, and thus the speed of travel of the spot horizontally across the screen.

Let it be supposed that a sample of the 50 Hz mains voltage is applied to the Y plates and that the 'scope's timebase is switched off, so that the beam is not being deflected horizontally. The spot on the screen will then follow the rise and fall in voltage from a maximum in the positive direction to a maximum in the

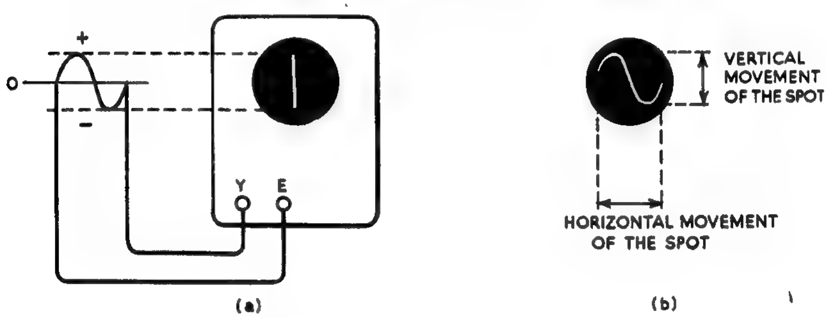


FIG. 9.3. *(a) With the timebase switched off, a vertical line corresponding in length to the peak-to-peak voltage of the applied waveform will be traced on the screen when the voltage is applied to the Y plates. (b) Showing how a sine wave is traced on the screen when the timebase repetition frequency equals the frequency of the applied signal.*

negative direction, as shown in Fig. 9.3 (a). A straight vertical line will thus be traced whose length corresponds to a value equal to the peak-to-peak voltage of the applied waveform.

Now, if under this condition the timebase is switched on so as to move the spot simultaneously across the screen in the horizontal direction, a waveform of a similar nature to that shown in Fig. 9.3 (b) will be traced on the screen. In order to produce one complete waveform of the 50 Hz mains voltage, however, it will be appreciated that the time taken for the spot to travel from left to right across the screen must equal the time of one complete sine wave. Thus, in the case illustrated, the repetition frequency of the timebase must be equal to the frequency of the mains voltage.

The foregoing discussion gives just a brief glimpse into the sphere of 'scope application. Additional details cannot be given here, but sufficient information has been given to explain how the 'scope can aid with alignment of f.m. receivers. Before returning to this subject, however, just a few words about the wobbulator.

When the 'scope is used for alignment purposes the aim is to have portrayed on the c.r.t. screen the response curve of the circuit section undergoing adjust-

ment, so that the curve can be altered by trimmer adjustments, while it is actually being viewed, to conform to the necessary shape and width.

The wobbulator makes this possible. This piece of equipment is essentially an r.f. oscillator which can be tuned over the f.m. band and the i.f. spectrum and which, instead of being amplitude modulated with an a.f. signal, as with service oscillators and signal generators, is frequency modulated with a signal derived either from the timebase of an oscilloscope, or from some other signal source as required by the application in hand.

When the 'scope's timebase signal is used the wobbulator serves to swing or "wobble" the frequency of the signal applied to the receiver in direct accordance with the horizontal sweep of the 'scope's timebase, so that the *deviation* of frequency about the nominal frequency, as registered on the tuning scale of the wobbulator, embraces the frequency spectrum (pass-band) of the circuits undergoing adjustment.

Many wobbulators used for this purpose employ a reactance valve to frequency-modulate the internally generated r.f. signal in synchronism with the 'scope's timebase. The valve is usually arranged to represent a virtual capacitance, and is connected in parallel with the tuned circuit of the wobbulator's r.f. oscillator, while the grid of the reactance valve is connected to the X-plate terminal on the 'scope, via a matching terminal on the wobbulator. This connexion thus causes the output frequency of the wobbulator to wobble to and fro (plus and minus) about its tuned frequency at a rate governed by the repetition frequency of the 'scope's timebase, while the deviation extent of the "wobbled" signal is dependent on the amplitude of the timebase signal.

In some wobbulators the internal r.f. oscillator may not be continuously variable over the entire frequency spectrum as with a.m. service oscillators and signal generators. Sometimes two oscillators and a mixer are used, one oscillator being tunable in the usual manner by the tuning dial and the other tuned to a fixed frequency and connected across the reactance valve. The two signals are mixed, and the signal of resultant frequency used for application to the receiver. In other instruments the injection of an r.f. voltage from a calibrated signal generator is demanded. This signal, together with the f.m. signal generated in the wobbulator, is fed to a mixer valve, also embodied in the wobbulator, and the difference frequency thus produced is used for injection to the set under adjustment.

THE IDEA OF 'SCOPE WOBBULATOR APPLICATION

The idea of the wobbulator 'scope set-up is for the 'scope to produce a vertical deflection whose magnitude is representative of the output voltage at the receiver's detector stage. In the so-called "non-visual" arrangement, it will be remembered, a voltmeter is used for this purpose. Thus, in order to secure a vertical deflection, the Y plates of the 'scope are connected across the detector output.

Now, as the spot travels across the screen of the c.r.t., so the frequency of the signal fed from the wobbulator to the receiver changes linearly over the pass-band of the tuned circuits, thereby causing the voltage output from the detector to vary according to the receiver's response characteristics. It is easy to realize,

therefore, that a frequency-versus-amplitude curve (response curve) will be displayed on the screen.

Since the output voltage from the detector may not be large enough to provide sufficient vertical deflection of the spot when connected direct to the Y plates, it is sometimes necessary to use an amplifier. This is contained inside the 'scope and is known as the Y amplifier.

As a means of permitting ready adjustment of the width of the curve portrayed on the screen of the c.r.t., most wobbulators embody a variable potentiometer connected between the X deflecting voltage and the grid of the reactance valve. Clearly, a too large X voltage will promote a frequency deviation extending well beyond the pass-band characteristics of the receiver and thus result in the display of a response curve too narrow to be of any useful purpose. Conversely, a too small X voltage will reveal only a portion of the curve, since the resulting frequency deviation will not embrace the full response width.

The repetition frequency of the timebase is not all so critical, but for the best results the frequency should not exceed about 50 Hz. Excessive flicker of the trace accompanied by distortion will result if too low a frequency is used.

If the displayed response curve appears inverted there is no need to worry unduly, since all the information relating to the response is available which ever way it occurs on the screen. The vertical deflection is relative to the potentials on the Y plates; thus, if a negative voltage, relative to chassis, occurs at the output of the detector, and this is applied direct to the Y plates, an upward deflection of the spot may occur. This will be reversed, and a downward deflection will occur, if a single-stage amplifier (that is, the Y amplifier) is introduced owing to the phase-change effect of the valve; the magnitude of deflection will also be increased as the result of the valve's amplification.

ALIGNING WITH A 'SCOPE AND WOBBULATOR

The method of connecting the apparatus for visual alignment of f.m. receivers is shown in Fig. 9.4. In order to assist with the alignment, a small "pip" can be produced on the displayed curve by injecting a marker signal from an accurately

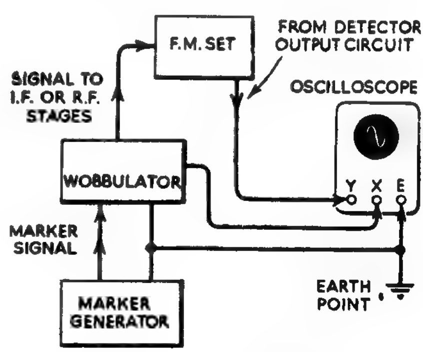


FIG. 9.4. *Method of connecting oscilloscope and wobbulator for aligning f.m. receivers. A marker generator aids in determining the width of response.*

calibrated signal generator to the receiver together with the wobbulator signal. The pip occurs on the curve at the frequency to which the marker generator is tuned; it thus becomes a simple matter to determine the overall width of response by running the marker pip from one side of the curve to the other and noting the

frequency range over which the marker generator has to be adjusted.

Most wobbulators have a terminal available for the injection of a marker signal, but if such a point is not available the marker signal can be loosely coupled to a suitable pick-up point in the receiver itself. Even if an injection point is available on the wobbulator, it is sometimes desirable to connect the marker signal direct to the receiver as a means of securing a pip of large enough size.

Nevertheless, in order to avoid distortion of the displayed response curve it is essential to maintain the smallest possible marker signal consistent with producing a marker pip of workable size. As the marker signal is increased the amplitude of the response curve will be seen to decrease, and as the signal is further increased the marker signal will tend to push the response curve out of shape. During the course of aligning, therefore, it is a good idea to switch off the marker generator occasionally as a means of ascertaining that the shape of the response curve is not in any way influenced by the marker signal.

Whether the Y voltage is picked up from the output of the detector or from a diode probe connected to the anode of the final i.f. valve is of little consequence. It it is decided to use the detector, however, the secondary of the discriminator transformer should be completely detuned so that the circuit will act as an ordinary a.m. detector, or rectifier, and produce a vertical deflection on the c.r.t. which is proportional to the gain of the i.f. stages.

ALIGNMENT PROCEDURES

As alignment procedures recommended by manufacturers will vary according to the circuitry employed in any particular receiver, the following details should be used only as a general guide.

The wobbulator signal, at the set's i.f. (nearly always 10·7 MHz), should be applied to the control grid of the i.f. stage prior to the discriminator, and the secondary and primary, in that order, of the associated transformer (usually the second i.f. transformer), carefully tuned to procure a response curve of a similar nature to that shown in Fig. 9.5 (a). The marker generator should be adjusted accurately to the nominal i.f., and the response curve should be resolved so that the marker pip falls at its centre frequency, as shown.

ALIGNMENT OF FIRST I.F. TRANSFORMER

Alignment of the first i.f. transformer should be carried out with the wobbulator signal applied to the grid of the associated stage (normally the f.m. mixer). Where this point is not readily accessible, the wobbulator signal may be capacitatively coupled to the mixer valve as was considered during the discussion of the non-visual method of alignment. The secondary and primary of the appropriate transformer should then be adjusted to produce a response curve similar to that shown in Fig. 9.5 (b). It will be observed that this is narrower than the curve at (a) owing to the inclusion of two more tuned circuits. When this curve has been obtained it should be flattened out by readjusting the core in the secondary winding until it looks something like that at Fig. 9.5 (c).

With the signal still applied to the mixer grid or i.f. test point on the tuner, the core in the primary winding of the discriminator transformer should be

adjusted to secure a curve similar to that at Fig. 9.5 (d). Finally, the core in the secondary winding, which was initially put out of adjustment, should be adjusted until a symmetrical double curve (Fig. 9.5 (e)) is obtained. Optimum symmetry

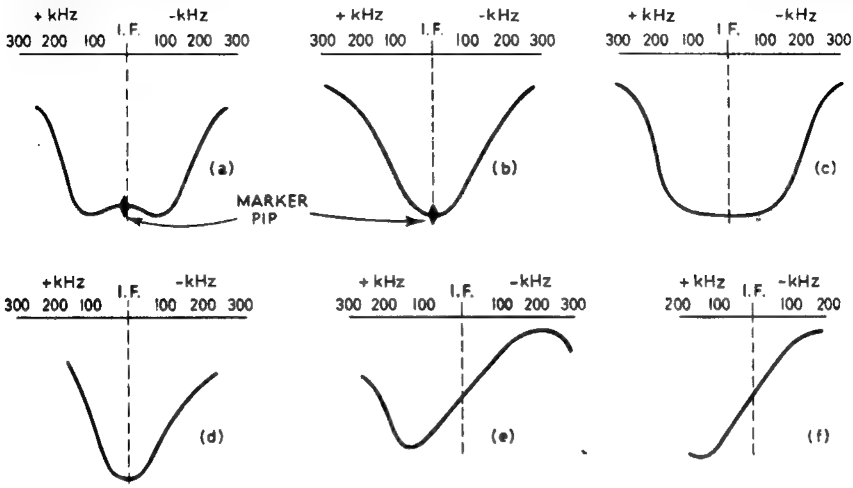


FIG. 9.5. *Response curves obtained when aligning the tuned circuits of an f.m. receiver, using an oscilloscope and wobbulator. Marker pips are shown on curves (a) and (b). The marker generator should be adjusted accurately to the nominal i.f., and the response curve should be resolved so that the marker pip falls at its centre frequency.*

of response occurs only when both diode circuits are in perfect balance, and it is as well to spend a little time over this final adjustment, checking and rechecking for balance and linearity under conditions of both strong and weak signal input.

It should be mentioned that if a diode probe was used for the production of a Y voltage during the alignment of the second and first i.f. transformers, the probe should be removed and the Y voltage picked up from the a.f. take-off point of the detector, usually across the volume control, when adjustment is made to the secondary of the discriminator transformer.

The above adjustments are best made with the wobbulator adjusted to give a sweep of approximately plus and minus 300 kHz. This is, of course, in excess of that required under normal operating conditions, but it does make the response curve of easily workable width.

The curve at Fig. 9.5 (f) should be obtained with the wobbulator adjusted for a sweep of plus and minus 200 kHz, which represents the approximate bandwidth required of a correctly aligned f.m. receiver. The oscillator circuit should be adjusted for accurate scale calibration, and the r.f. circuits for maximum output; the principles detailed in the section on non-visual alignment also apply in this connexion.

It is most important to avoid overloading the receiver or Y amplifier, if used, by applying the smallest possible signal consistent, of course, with sufficient vertical deflection of the trace. The result of overloading will produce a curve with a very much flattened top which, unless one is well versed with this mode

of alignment, will almost certainly give the impression that the receiver is ideally aligned.

With combined a.m./f.m. receivers, the two alignment processes are totally independent of each other and, unless the makers' instructions stipulate the order in which they should be aligned, it matters little which section first receives attention. Alignment of the a.m. section follows normal practice.

TRANSISTOR RECEIVERS

Exactly the same alignment procedures are adopted with transistor tuners as detailed above for valve receivers. Fig. 9.6 shows the best way of connecting the output of an f.m. generator or wobbulator to the input of an i.f. stage. This

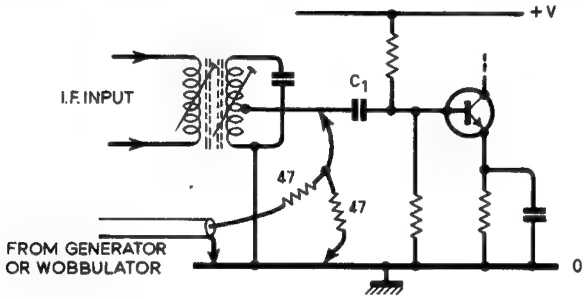


FIG. 9.6. *Method of connecting generator output to input of transistor i.f. amplifier.*

scheme keeps the generator loading pretty constant, and the injected signal strength is about 10 per cent of the terminated feeder signal voltage.

The generator must be d.c.-isolated from the base circuit of the transistor to which the signal is applied. In Fig. 9.6, C1 gives the necessary isolation; but in circuits where there is no such capacitor, a capacitor must be connected in series with the "live" signal generator connection. A capacitor about 0·01 μF is suitable. Without this isolation the base circuit could be shunted by the low resistance of the generator's attenuator.

STEREO RECEIVERS

As indicated in Chapter 7, the accuracy of the i.f. alignment of stereo receivers to some extent governs the crosstalk performance of the decoder. Insufficient

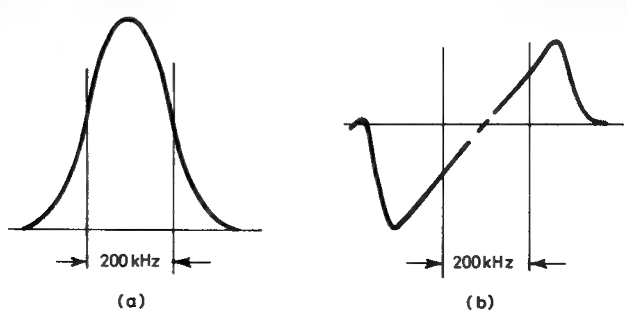


FIG. 9.7. *I.f. (a) and discriminator (b) responses obtained from stereo tuner (see text).*

bandwidth or wide bandwidth with phase distortion can seriously impair the separation of the A and B channels.

Fig. 9.7 shows the type of displays that should be aimed at in stereo equipment. Displays like this were obtained from an i.c. f.m. tuner with the generator input capacitively coupled to the output of the mixer (collector of bipolar or drain of f.e.t.) and with the 'scope Y-input picked up from the diode circuit shown in Fig. 9.8 connected to the input of the final i.c. in the i.f. channel (e.g., that feeding

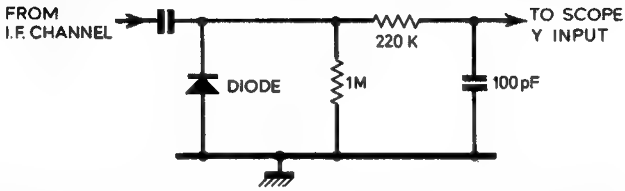


FIG. 9.8. *Diode system used for obtaining the response at (a) in Fig. 9.7.*

the f.m. detector transformer) to give display (a) and with the 'scope Y-input connected to the audio output from the f.m. detector direct (without the diode circuit) to give display (b). The tuning cores were adjusted to give symmetry and maximum amplitude of (a), with a bandwidth of 200 kHz between the —3 dB points, and high linearity over 200 kHz at (b). If the 200 kHz swing approaches non-linear regions on (b) crosstalk and adjacent channel "birdies" are likely to be troublesome (see Chapter 7).

Details for aligning the multiplex decoder are given in Chapter 7.

CHAPTER 10

Servicing F.M. Receivers

PROVIDED due consideration is given to the higher signal and intermediate frequencies, servicing domestic f.m. receivers, adaptors and combined a.m./f.m. receivers is little different from servicing ordinary a.m.-only receivers of modern styling. Valve equipment is first dealt with, followed by transistor equipment on page 187. As with a.m. equipment, analysing for total failure—and for certain faults—is best done by dividing the whole circuit into six sections and then concentrating on the individual sections in the following order: (1) the circuits concerned with the production and transmission of the h.t. and heater supplies; (2) the a.f. circuits; (3) the f.m. detector (discriminator); (4) the i.f. stages (including the limiter if fitted); (5) the frequency-changer stage (mixer and oscillator); (6) the r.f. stage.

In Fig. 10.1 is given a circuit diagram of a representative f.m.-only receiver of commercial design. Let us first run through this as a means of tying up with what has already been discussed. Valve V1 provides the function of r.f. amplification and frequency changing—section A being the r.f. amplifier and section B the self-oscillating frequency changer. It will be observed that the r.f. section is arranged in the earthed-grid mode, and that the aerial signal is applied via L1 and L2 to the cathode circuit. The amplified signal is thus developed in the anode circuit across L3. This signal is applied to the frequency-changer stage at the point of minimum oscillator voltage which, it will be remembered, exists at the junction of the two 6 pF capacitors when the bridge-type oscillator circuit is properly balanced. Coil L4 and trimmer T1 form the oscillator tuned circuit which adopts variable permeability tuning in gang with L3. Coil L5 serves to provide the correct degree of positive feedback necessary for oscillation.

The i.f. signal appearing in the anode circuit of V1B is developed across the first i.f. transformer and conveyed to the pentode valve V2 for amplification. Further i.f. amplification is given by V3 and the signal is raised to the level necessary for proper operation of the ratio detector, comprising the discriminator transformer, diodes V5A and V5B and associated components. An interesting feature here is that the d.c. potential developed across the 33 k load resistor and 8 μF stabilizing capacitor is fed back as bias to the *suppressor grid* of V3 giving automatic gain control. The tuning indicator V4 is also fed from this circuit via a step-down potential divider comprising the 2·2 megohm and 1 megohm resistors across the ratio detector load.

The a.f. signal is developed across the 100 pF capacitor connected between

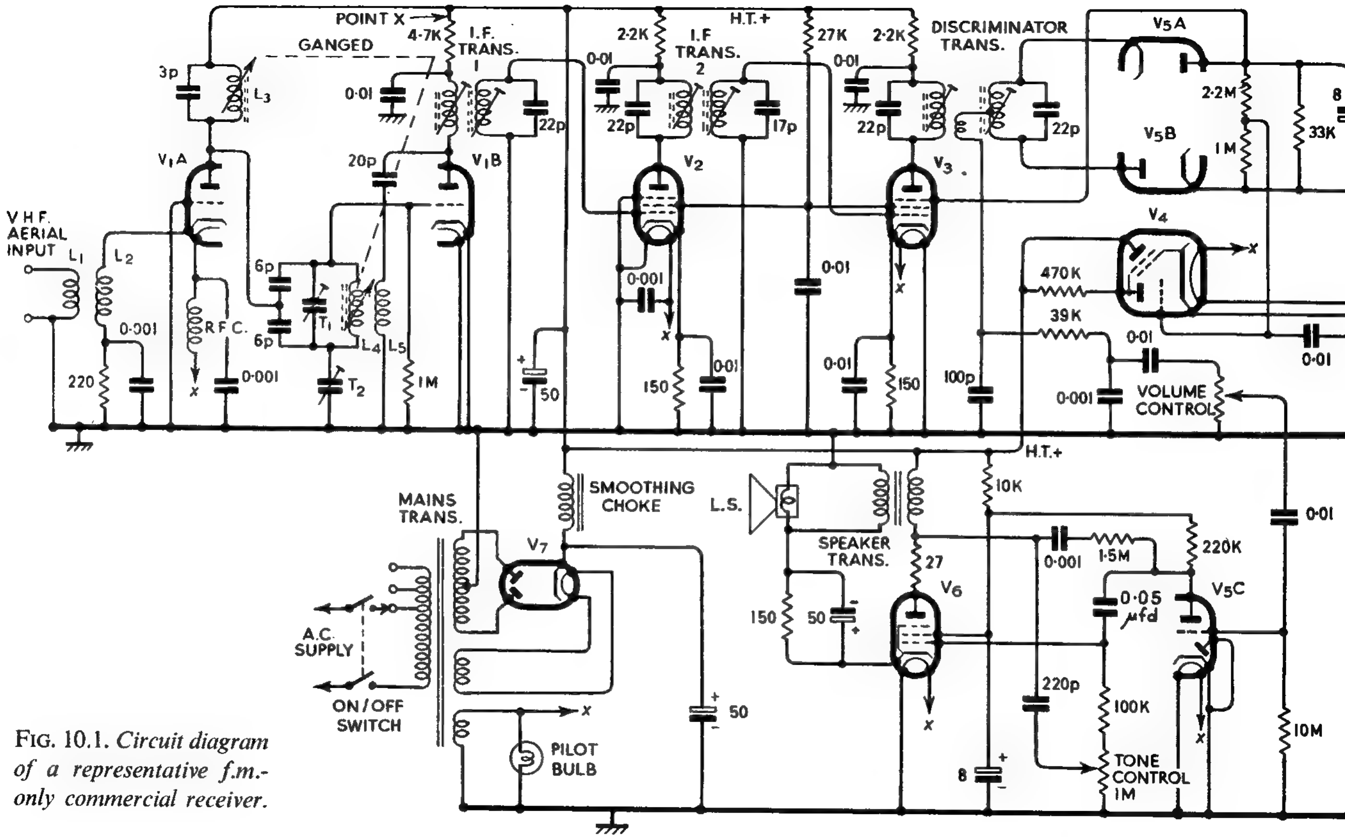


FIG. 10.1. *Circuit diagram of a representative f.m.-only commercial receiver.*

the tertiary winding and chassis. The 39 k resistor and 0·001 μF capacitor provide de-emphasis and the signal is coupled to the volume control by way of the 0·01 μF capacitor. This allows the passage of the desired amount of a.f. for amplification by V5C. The amplified a.f. signal is developed across the 220 k resistor in the anode circuit of V5C and is conveyed, via the 0·05 μF coupling capacitor, to the control grid of the output valve V6.

A degree of negative feedback is given in virtue of the coupling between the secondary of the speaker transformer and the cathode of V6. This tends to even the a.f. response and reduce the distortion content. Negative feedback tone control is also provided by the 1-megohm tone-control potentionmeter and the 220 pF capacitor connected to its slider.

High-tension is supplied by the indirectly-heated full-wave rectifying valve V7. A mains transformer is used to feed the rectifying valve and also to provide low-tension for the valve heaters. H.T. smoothing is carried out by the smoothing choke and the two 50 μF electrolytic capacitors in the conventional manner.

SET COMPLETELY DEAD

Total failure of the receiver would lead one first to check on the mains supply. Trouble here would most likely be responsible if the pilot bulb and valve heaters were not alight. Other possibilities are (1) a break in the receiver's mains lead; (2) a fault in one section of the mains on/off switch (this is invariably ganged either to the volume or tone control); (3) open circuit of the primary winding of the mains transformer or a poor electrical connexion of the mains-voltage selection switch.

If, on the other hand, the pilot bulb and valves are glowing normally, it will be necessary to delve a little deeper, firstly into the circuits concerned with the production of the h.t. supply. The fact that the valves are alight is usually sufficient indication that the l.t. circuits are in order, and no tests are warranted in this connexion.

Lack of h.t. voltage between the cathode of the h.t. rectifier and chassis, accompanied by a cold rectifier, will almost certainly mean that the emission of the rectifier has failed. If the valve is glowing red and operating very hot, then a short-circuit in one of the electrolytic smoothing capacitors should be suspected. The receiver should immediately be switched off in an endeavour to save the valve and a test made for a short on the h.t. line with respect to chassis.

If the h.t. line voltage is normal, or a little above normal, one can usually discover quickly whether the output stage is working by placing an ear close to the loudspeaker and listening for a trace of residual mains hum. If a slight hum is present and the output valve is fairly warm, the output stage is probably in order and this can be proved by touching the control grid of the valve with the blade of a screwdriver. Normal operation will promote a considerable rise in hum level when the grid is touched.

Voltage checks on the anode and screen electrodes of the output valve will be called for if the output stage is proved defective. Often the primary winding of the loudspeaker transformer becomes open-circuited, or a screen feed resistor fails. A short-circuit in a capacitor in the anode circuit of the valve is another cause of the trouble, but this will be revealed by lack of voltage on the valve's

anode and possible over-heating of the speaker transformer. An open-circuit cathode resistor would, of course, cause the same symptom, but resistors in this position are somewhat less prone to failure. The valve itself must also be considered, and if it is cool, even though its heater is glowing, lack of emission should be suspected. If the screen grid of the valve is glowing red, the trouble is almost certainly due to failure of the primary winding of the speaker transformer.

Once the output stage is proved in order, the next simple move would be to touch the grid of the a.f. amplifier valve with the blade of a screwdriver while actually holding the blade in the hand. If the other hand is kept in the trouser pocket there will be no danger of shock. Goodness of the a.f. stage will be evidenced by a very loud hum from the loudspeaker. Should this not happen, then the valve should come under suspicion. Lack of anode voltage due to failure of the anode load resistor is a frequent cause of the trouble.

We have now reached the volume control and have covered the first and second sections as detailed at the beginning of the chapter. Next in order, assuming that the fault still exists, comes the f.m. discriminator section.

The best method of checking this section, and the i.f. stage prior to the discriminator, is by injecting an unmodulated signal at i.f. at the control grid of the final i.f. valve and, in the case of a ratio detector, observing the effect on a meter connected across the stabilizing capacitor. The instrument connexions are the same as for alignment, which was dealt with in the previous chapter. If the final i.f. stage and discriminator are working properly the meter will show a deflection, provided sufficient signal is applied to the i.f. valve, which will vary as the amplitude of the signal is slowly varied. Should the final i.f. valve be arranged also as a limiter, however, an excessive signal will promote limiting action and will detract from the goodness of this test.

A Foster-Seeley discriminator can be checked in a similar manner, but here a sensitive meter should be connected in series with a resistor of 47 k ohms, the precise function of the resistor being to isolate the lead capacitance from the circuit under test, and the meter/resistor combination connected across both load resistors. If the section is working properly equal but opposite readings should be obtained on the meter over a frequency band of approximately plus and minus 100 kHz. At the nominal i.f. the meter should indicate zero voltage.

Most failures of discriminators are due to faulty valves; failure of one diode will preclude efficient operation on f.m., though a weak distorted signal might be received slightly removed from the proper setting on the tuning scale when the signal falls on the sloping side of the response due to one diode. The effect would be much after the style of f.m. reception secured on an a.m. receiver by the detuning artifice.

At this stage, one should not be tempted to alter the tuning of the i.f. transformers unless, of course, it is obvious that the trimmers have been disturbed by a former operator. In this case the entire alignment procedure should be followed before making any further tests.

The remaining i.f. stages can now be checked by leaving the meter connected to the discriminator and transferring the signal generator output in turn to the control grids of the previous valves. Testing in this way should be carried out as far as the signal grid of the frequency changer valve, turning down the output

of the signal generator as the other valves contribute to the overall sensitivity of the i.f. channel. The input signal should be progressively reduced to maintain much the same meter deflection as before. It is important to avoid overloading, as the resulting damping of the tuned circuits due to grid current tends to widen the i.f. pass-band and make accurate tests impossible.

A fault in any i.f. stage will be revealed by a decrease in meter deflection when changing from one i.f. stage to the preceding one. This is usually caused either by valve trouble or by open-circuit of an anode, screen or cathode resistor. If the anode and screen voltages are normal but no voltage can be measured across the cathode resistor—positive at the cathode with respect to chassis—then the valve is almost certainly low in emission.

We have now arrived at the frequency-changer section, and if no broadcast programme can be heard when the oscillator is tuned over its range, the trouble exists somewhere in the v.h.f. tuner. Substituting the valve with one of known goodness represents the next most logical step to take. If the set is still dead, a check should be made to see that the oscillator is functioning. This is best done by inserting a milliammeter in series with the lead from the h.t. line to the anode of the valve, at point X marked on the circuit in Fig. 10.1, and then short-circuiting the grid of the oscillator valve to chassis, preferably via an 0·1 μF capacitor. A change in current reading will indicate oscillation. If this does not happen, the associated resistor and capacitors should be checked for goodness; lack of current indication to start with, of course, will either indicate that the valve is faulty or that the anode resistor (or possibly the primary of the first i.f. transformer) is open-circuited.

Apart from the valve and open-circuit of the r.f. coil, there is nothing that can cause much trouble in the r.f. section. Not very often does the cathode resistor give trouble here.

Having progressed so far and still signals are not receivable, the trouble is almost certainly caused by misadjustment of the oscillator. A fairly strong signal from a signal generator, tuned to the centre of Band II, should be applied to the aerial terminals of the receiver, and the oscillator tuning should be adjusted for maximum indication on the meter, after roughly setting the receiver tuning to the centre of its range. Final adjustments of the oscillator and r.f. sections should be carried out as detailed in the previous chapter. Open-circuit or value deviation of one of the capacitors associated with the tuning of the oscillator might prevent correct adjustment of the oscillator trimmer in relation to Band II.

GUIDES

Some of the foregoing information does not apply solely to f.m. receivers, since it might equally be of use during the course of servicing a.m.-only models. Its inclusion, however, was considered desirable for the sake of completeness in this chapter.

The loudspeaker output is a good guide during the servicing process. No signals but a noisy background at full volume setting should lead one first to suspect failure of the local oscillator. Lack of signals and noise but a slight hum from the loudspeaker suggest that an i.f. stage is defective.

In receivers featuring a tuning indicator, lack of signals from the loudspeaker

but normal operation of the tuning indicator indicate trouble in the a.f. stages.

The temperature of valves often indicates whether they are passing insufficient or excessive current. Similarly, an overheated resistor is a useful guide which nearly always indicates that an associated decoupling capacitor is short-circuited.

COMBINED RECEIVERS

In Fig. 10.2 is given a circuit diagram of a representative a.m./f.m. receiver. With such receivers servicing is aided by observing whether the set is dead or faulty on both services. For instance, if the receiver is normal on a.m. but dead on f.m., the trouble would most likely lie in the v.h.f. tuner itself; the discriminator or the f.m. intermediate-frequency transformers could be responsible, of course, but these sections seem to give less trouble. Routine testing should be carried out along the lines already described.

In the receiver in Fig. 10.2, valve V1 is concerned solely with the f.m. function, valve V2 serves on both services; on f.m. the heptode acts as an i.f. amplifier and the triode a.m. oscillator is shorted out by switch S11. Valve V3 is also common to both systems. The triple-diode-triode valve V4 has sections B and C operating on f.m. (in the ratio-detector circuit), section D (acting as a.f. amplifier) operating on both a.m. and f.m. and section A (acting as a.m. detector) operating on a.m. only. The output valve V6 and the rectifying valve V7 are also used on both services. The magic-eye tuning indicator V5 is operational on both a.m. and f.m.

The switching is somewhat complex and trouble here might give rise to a fault on one service. For example, the failure of S6 in coming into contact on f.m. would prevent h.t. from reaching the v.h.f. tuner, while having no effect at all on a.m.; S11 remaining in contact on a.m. would mute the a.m. oscillator, and while giving operation on f.m. would prevent it on a.m.

S3 switches in and out the v.h.f. tuner and S4 provides a similar function in regard to the a.m. aerial-tuned circuits. S7 switches out the a.m. radio-frequency section of the tuning gang on f.m. S8 short-circuits the primary of the second f.m. intermediate-frequency transformer on a.m. S14 cuts off h.t. to the a.m. oscillator on f.m. S15 selects the a.m. detector, while S16 selects the ratio detector on f.m. The other switches are those conventionally used on a.m.-only receivers for band changing and for switching in a gramophone pick-up.

An extra tuned circuit, L3 and the 47 pF capacitor, is incorporated in the v.h.f. radio-frequency stage. This serves as an i.f. filter and should be tuned for minimum output when a strong signal at the intermediate frequency is applied to the aerial terminals. A similar filter, L7 and the 1,500 pF capacitor, is used in the a.m. aerial input circuit. This, of course, is tuned at the a.m. intermediate frequency.

Trimmer T2 on both circuits in Figs. 10.1 and 10.2 serves to balance the oscillator circuit so that the point of coupling to the r.f. stage is at minimum oscillator voltage. The method of adjusting this is dealt with in Chapter 4.

CHECKING FOR DISTORTION

Distortion in an f.m. receiver is nearly always due to misalignment or disturbed balance of the discriminator. Full details for checking this section

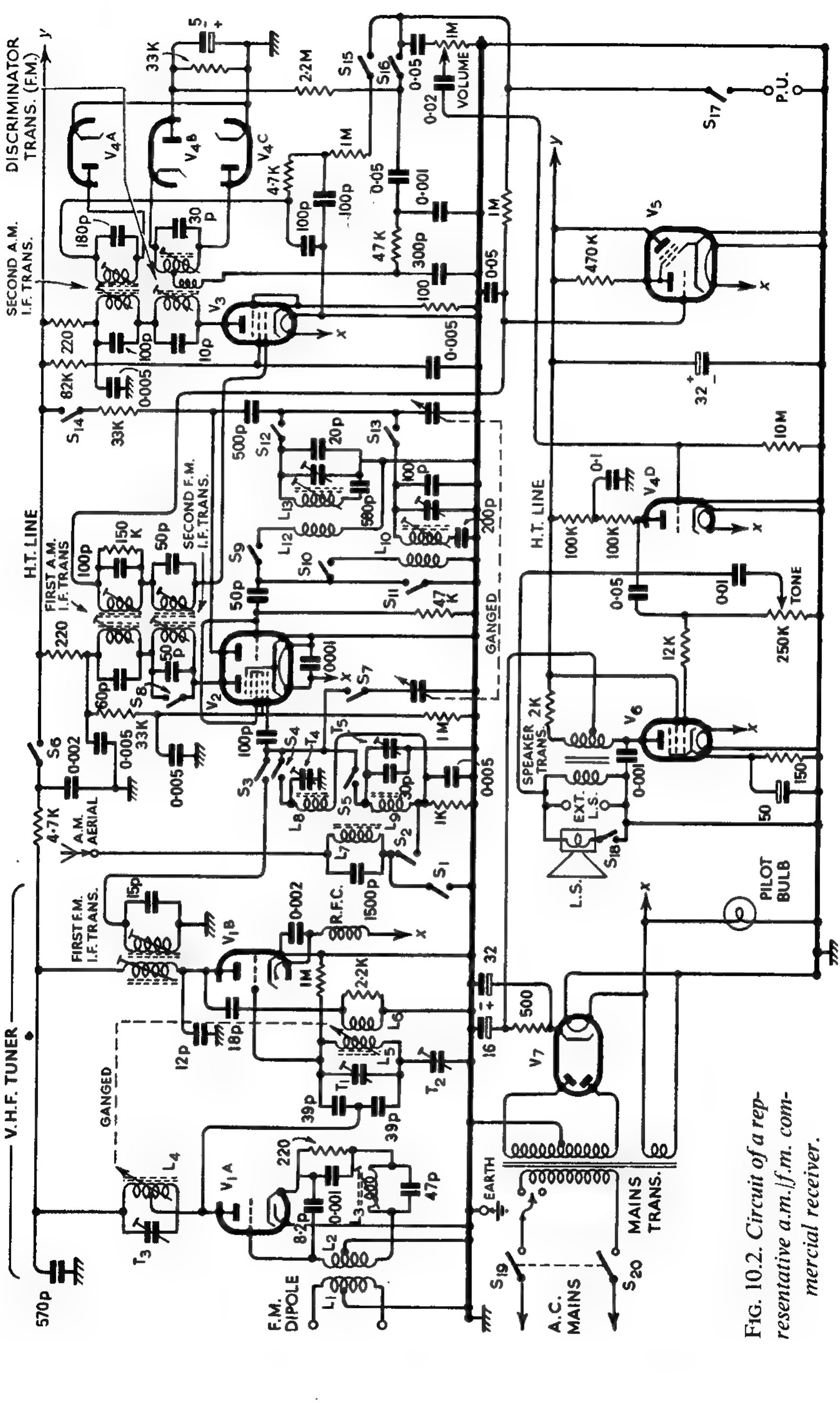


FIG. 10.2. *Circuit of a representative a.m./f.m. commercial receiver.*

have already been given. If tests reveal that the diode circuit is out of balance, and if normal adjustments of the discriminator transformer do not help matters, the emission of one diode with respect to the other may have altered. With receivers featuring double-diode and triple-diode-triode valves replacement of the valve is usually necessary to prove this trouble.

Where two semiconductor diodes are employed, severe lack of efficiency of one diode can often be proved by comparing the forward and reverse resistances of the diodes. If both diodes are in order they should possess near enough the same values. As was previously intimated, only very rough matching of diodes can be achieved by resistance measurements. What is really needed is a method whereby a comparison of dynamic resistance at the working point of the diodes can be made.

In practice, however, this is not a simple matter because the "working point" of the diodes in a ratio detector, for example, depends on the strength of the i.f. signal applied to the circuit. This might well vary between programmes so that, although good matching is achieved on, say Radio 3, mismatching may occur on Radio 2 or Radio 4. If a limiter stage is used prior to the ratio detector this trouble is dispelled to a large degree, always provided that the signal strength corresponding to each programme is high enough to promote limiting.

Observing the discriminator response on a 'scope gives a true picture of the balance of the circuit under varying input voltages and will immediately reveal whether the efficiency of one diode circuit is below or above the efficiency of the other. In balanced circuits there is always a possibility of one load resistor changing in value and unbalancing the network.

If a 'scope is not available, matching of germanium diodes can be achieved by applying an a.m. intermediate-frequency signal to the control grid of the last i.f. valve and metering the a.f. output. The signal generator should be adjusted to the nominal i.f. and then accurately adjusted for minimum a.f. output. Without disturbing the tuning of the signal generator, the diodes should be interchanged and if they are matched, the frequency of the applied signal will still correspond to minimum a.f. output. If, on the other hand, it is found necessary to alter the frequency of the input signal slightly to secure minimum a.f. output, the diodes are probably mismatched. Other diodes should be tried in turn until a pair is obtained which, when interchanged, necessitates the smallest possible change of input frequency to maintain minimum a.f. output.

When making this test on a ratio-detector circuit it is best to remove the stabilizing capacitor. After conclusion of the test, the capacitor should be reconnected and the discriminator transformer adjusted for optimum balance in the usual manner.

Distortion may creep in as the receiver warms up. This is nearly always due to drift of the oscillator and can be corrected by retuning. Unfortunately, this symptom in a mild form occurs on a number of receivers and tuners during the initial warming-up period. The effect in excess, however, should lead to investigation of the oscillator circuit as a whole. An ageing oscillator valve, alteration in value of a capacitance in the oscillator stage, overheating of the stage due to insufficient ventilation or to the use of a wrong mains-voltage tapping, looseness of the oscillator coil on its former and movement of the oscillator coil tuning

slug due to vibration from the loudspeaker represent the chief causes of the trouble.

Normal a.f. distortion, although being of a similar kind to that produced by discriminator trouble and misalignment, requires an entirely different approach. Usually the cause lies in the output stage. A low-emission valve and incorrect bias on the output valve are frequent causes. An electrical leak in the a.f. coupling capacitor, connected between the anode of the a.f. amplifier and the control grid of the output valve, causes the output-valve grid to go positive with respect to cathode and thus counteracts the standing negative bias. This fault promotes considerable distortion and overheating of the output valve. Distortion is also produced by an increase in value of the anode load resistor of the a.f. valve; that is, the 220 k resistor in the anode circuit V5C in Fig. 10.1, and the 100 k resistor in the anode circuit of V4D in Fig. 10.2.

A loud hum at all settings of the volume control is almost certainly the result of failure of one of the electrolytic smoothing capacitors associated with the h.t. rectifying valve. Hum which is present whether or not the receiver is tuned to a station, but which decreases when the volume control is turned down, is probably occurring in the a.f. stage. A leak between the heater and cathode of the valve is a possible cause.

If the hum is present only when the receiver is tuned to a station, then the trouble lies either in the discriminator valve or in the stages prior to the discriminator. A rather interesting hum which was investigated by the author occurred only when the combined a.m./f.m. receiver was accurately tuned to an f.m. transmission. Slight detuning, within the range of the i.f. and discriminator response, caused the hum to disappear. Thus, if the receiver was accurately tuned, the hum was present but would soon disappear as the receiver warmed up and the oscillator drifted slightly. Conversely, if the receiver was tuned less carefully, the hum would not be present initially but would occur as the tuning drifted to the critical point. It would last for a while and would then cut out as the tuning drifted past the critical point. The trouble was eventually traced to a small heater-to-cathode leak in one of the diode sections of the triple-diode-triode valve.

A similar manifestation of hum occurs as the result of hum-frequency modulation of the oscillator valve from its h.t. or heater supply. Insufficient filtering of the h.t. feed should be suspected and, if necessary, extra smoothing, possibly in the form of a larger electrolytic capacitor, should be given to the h.t. supply. If the trouble is in the heater supply, it can usually be cured by decoupling the heater pins to chassis via 100 pF capacitors. A faulty oscillator valve may cause trouble, and in persistent cases it is desirable to try a number of valves and select the one less susceptible to the effect.

As a means of avoiding undesirable couplings in the r.f., frequency-changer and i.f. stages of v.h.f. receivers, chassis connexions relating to an individual stage are made at a common earthing point as shown on the circuit in Fig. 10.3. It is most important, therefore, to ensure that replacement parts are routed as in the original layout and earth connexions made to the original chassis points. It is as well also to connect the components into the circuit with lead lengths as the originals which, in a number of cases but not all, are as short as possible.

It is interesting to note that little over two inches of wire close to the chassis can effectively by-pass signal frequencies of the order of 100 MHz.

Incorrect placement or earthing of new components might well lead to instability, detuning of circuits, reduced sensitivity and radiation from the

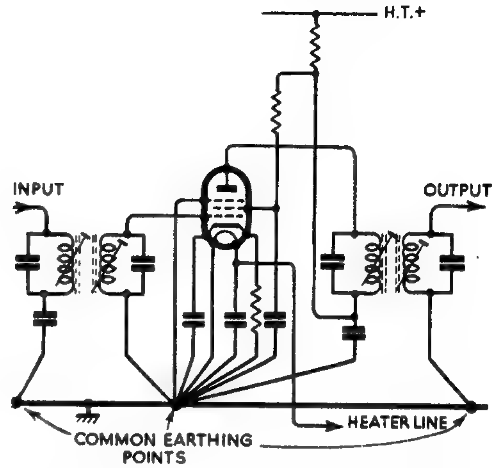


FIG. 10.3 *How the chassis connexions relating to an individual stage are made at a common earthing point.*

oscillator to the aerial circuits, to name only a few of the possible adverse disturbances.

In this respect, screening cans and covers should also be securely fitted after any servicing operation demanding their removal. In some receivers a half-wave dipole, loaded with an inductor, is fitted inside the cabinet. Due to the cabinet dimensions, the required length of wire cannot usually be accommodated, so the aerial is tuned with an inductor at the centre and the required matching impedance tapped off along the inductor. Feedback between this internal aerial and the tuned circuits in the receiver sometimes occurs if the screening cans and covers are not making one-hundred-per-cent electrical contact with the receiver chassis. Moreover, as the chassis in a small table receiver can affect the characteristics of the internal aerial, adjustments to the r.f. section are best made with the receiver completely assembled in the cabinet.

Experience has shown that from the f.m. aspect, the v.h.f. tuner tends to give most trouble. Often a badly soldered connexion or poor connexions between the valve pins and valve-holder sockets gives intermittent operation, or detuning, on f.m. which is clearly evidenced by subjecting the chassis to a sharp knock.

TRANSISTOR EQUIPMENT

The same general principles apply to the servicing of transistor equipment, but, of course, the current working of the bipolar transistor has to be taken into account. A valve calls for signal *voltage* at its control grid, giving it a high input impedance, while a bipolar transistor works by signal *current* swings in the emitter/base junction, giving it a low input impedance. F.E.T.s are different from bipolars in this respect for they, like a valve grid, require signal voltage at the gate electrode. This is because the input signal controls current through a

junction indirectly by means of an electric field. It thus has a high input impedance as a valve.

I.c.s are complete "packets" or microcircuits containing a number of transistor elements, diodes and resistors, often with some of the diodes acting as capacitors by being reverse-biased. It is best to regard an i.c. as an entity when servicing, rather than looking upon it as an item of equipment containing separate components. The i.c.s used in domestic equipment represent "operational amplifiers" of linear mode (as distinct from the digital i.c.s of computers) and it is desirable to consider them merely as amplifying "units" in the equipment as a whole.

How the various elements are arranged within the encapsulation relative to the circuit as a whole depends on the requirements. Sometimes all the transistor elements are geared into a single amplifier system, while other designs may call for two or more amplifier channels within the one i.c. The way that the lead-out wires are connected in circuit gives a clue as to how the i.c. is tailored into the circuit as a whole.

I.c.s have a value of open-circuit gain without feedback and this is often very high indeed (60 dB or more). The gain is tamed for the requirements in hand by the application of negative feedback. Similarly, the device has an overall frequency response which can also be modified by feedback. In action, though, the gain is "channelled" into the required frequency spectrum by filters and tuned circuits. This, of course, is just the same with transistors, for these can have a very wide inherent bandwidth, the working spectrum being controlled by tuned transformers, filters and so on.

As indicated in the circuits of some of the earlier chapters, i.c.s are represented by large arrow heads, pointing in the direction of the signal through the equipment. The circuits of earlier chapters also show the symbols of bipolar and f.e.t. devices.

The general pattern of a circuit is the same whether valves or transistors are used as the active devices; the main difference lies in the values of the passive components. Fig. 10.4 gives the circuit of an f.m./a.m. transistorized receiver minus the audio stages (for these refer to my *Hi-Fi and Tape Recorder Handbook*: also see Chapter 11). The various stages are marked on the diagram, as also are the voltages which would be expected at the various electrodes of the transistors under a.m. and f.m. working conditions.

Electrode voltages give good clues as to how the stages are working, but for the low voltages to be measured correctly without the operation of the stages being affected unduly the testmeter should have a high sensitivity (about 20,000 ohms/volt) and incorporate low-voltage full-scale ranges.

When injecting signals into a transistor receiver for r.f., i.f. or a.f. testing, the "live" output from the signal generator should be isolated from the transistor circuit by a suitable value capacitor to avoid the generator's attenuator from shorting the bias of the transistors, an action that could precipitate transistor failure.

The majority of contemporary receivers are designed on printed circuit boards or "modules", making it difficult to remove active and passive components for substitution tests. However, it is often possible to discover how a bias feed circuit

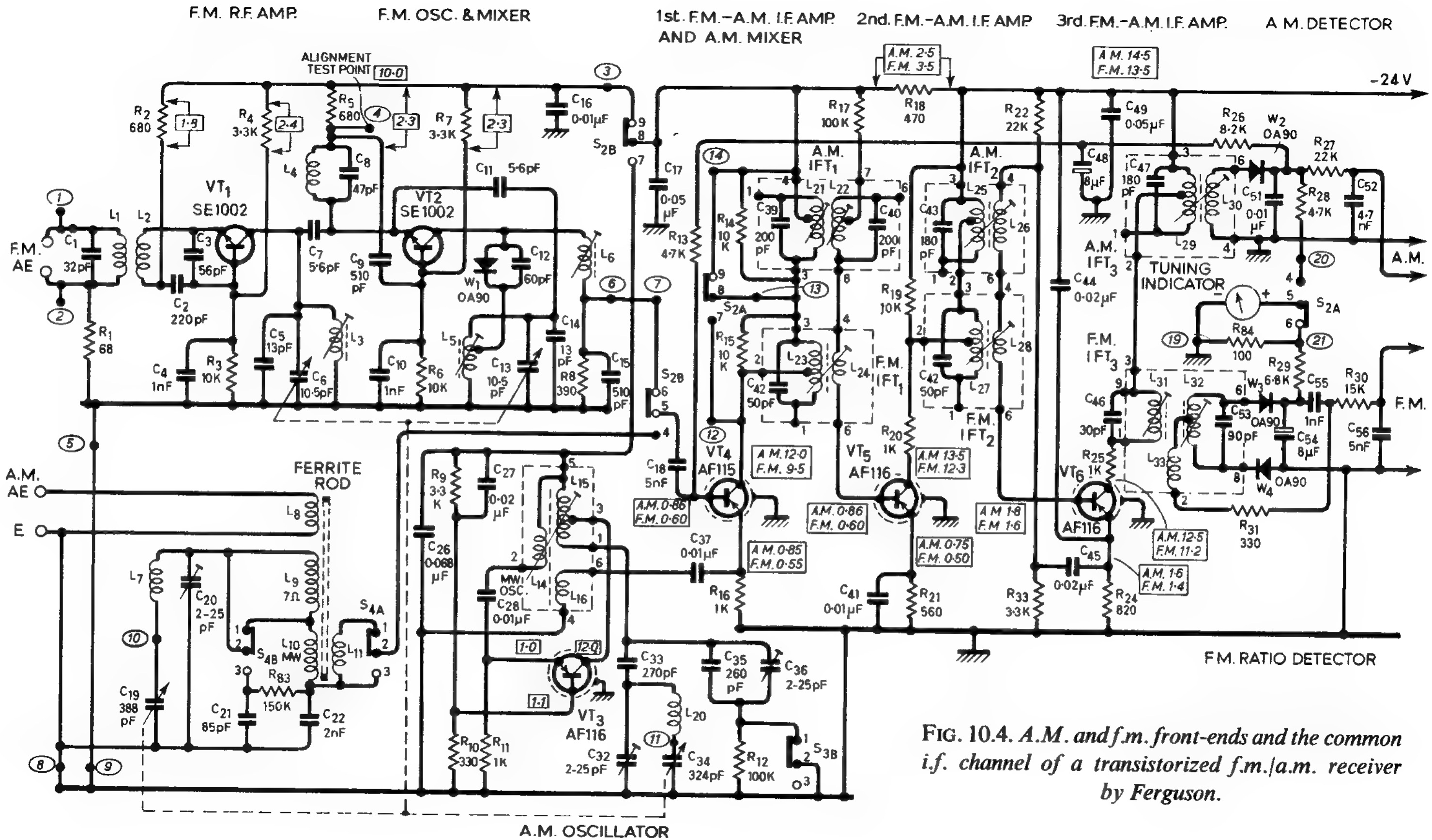


FIG. 10.4. *A.M. and f.m. front-ends and the common i.f. channel of a transistorized f.m./a.m. receiver by Ferguson.*

or transistor is operating by measuring the voltage developed across, say, an emitter or collector resistor and translating this, relative to the value of the resistance, into current. For example, 10 V across a 1 k resistor indicates a current of 10 mA. The working of a transistor, too, can be revealed by measuring the voltage across a collector or emitter resistor and then slightly adjusting the base bias by shunting extra resistance across an arm of the potential divider while observing any change in voltage reading. One must be careful, though, not to over-run the base, so to speak, for this would cause a high value of collector current and possibly destroy the emitter/base junction. It is best to shunt the lower arm of a base potential divider to *reduce*, rather than increase, the base current. This would then cause a drop in emitter or collector current, and hence a corresponding drop in measured voltage across the resistor, of a correctly working stage. All this can be done, of course, without removing any components from the printed circuit board.

I.c.s are less accommodating in this respect, and if such a component is suspect a replacement generally has to be tried, making sure beforehand that a short in the circuit was not responsible for the failure, otherwise the replacement would suffer a similar fate. In practice, quite a bit of information on the operation of an i.c. stage (or transistor stage for that matter) can be gleaned by the signal injecting technique, starting towards the detector end of the circuit and then moving stage-by-stage backwards towards the aerial input until the signal ceases to pass through the circuit, at which point the trouble would exist.

In Fig. 10.4, for example, when the symptom is failure of both f.m. and a.m., an f.m. or a.m. signal could be applied to the base of VT6 and the signal monitored at the appropriate detector ouput. If response is obtained at that base but not at the base of VT5, say, then the trouble would exist between the input of VT5 and the base of VT6.

If only one service is dead, however—say, f.m.—then one could be sure that the i.f. channel is active since a.m. passes through the common i.f. channel. In this case the trouble would lie in the f.m. front-end section.

Short-term frequency drift is not uncommon in transistor equipment. This is often caused by the transistors in the front-end becoming thermally stable. However, the drift should not be excessive or in advance of the pull-in range of the a.f.c.

For transistor servicing information in greater detail the reader is referred to my *Rapid Servicing of Transistor Equipment* and *Radio and Audio Servicing Handbook*.

Adjustments etc. in stereo decoders are considered in Chapter 7.

CHAPTER 11

Equipment and Specifications

DURING the time of writing this second edition an outstanding trend geared to f.m. reception has been the tuner-amplifier. Hosts of music lovers and hi-fi enthusiasts are making this compact unit the nucleus of a high quality sound reproducing system. A tuner-amplifier is an integration of a full-power stereo amplifier and a tuner unit, made possible in a small volume by solid state electronics. The tuner side is always designed for receiving the f.m. transmissions in stereo as well as mono, and even though a stereo decoder may not be fitted at the time of the purchase the tuner section will feature facilities allowing a decoder "module" to be easily plugged in at a later date when the owner is lucky enough to be served by a local stereo-encoded station. Many tuner-amplifiers, however, are stereo-ready which makes it possible for the owner to try his hand at long-distance reception from a stereo-encoded transmitter removed from his normal range of reception (see Chapter 8).

The tuner might also be equipped to receive a.m. transmissions; either merely the medium waveband or the medium, long and short wavebands. The reason why the lesser quality a.m. bands are sometimes served in addition to f.m. is because a purchaser of a tuner-amplifier is not uncommonly expecting the new equipment to replace an old radio or radiogram in terms of programme potential, so if he has been a keen listener of the "popular" programmes at the high frequency end of the medium waveband he will need this waveband at least on the new equipment. Similarly, if he has been a keen short-wave listener.

However, for those listeners whose music tastes are satisfied by Radio 2 type of music and above, and who are not particularly keen to explore the "wild" short waves on a hi-fi set-up, then the f.m. band alone will suffice. It is only this band, of course, that can yield hi-fi quality reception and stereo-by-radio.

It is often wondered whether tuner-amplifiers are compromise items of equipment in view of the tuner/amplifier integration. They are certainly not, and in most models the tuner section is equally as good as a tuner designed individually; and the same applies to the amplifier part. The cost of a tuner-amplifier is generally slightly less than the two items of equipment separately because it is possible to make some parts common to both, like the power supply section, housing, main chassis work and so forth.

The trend is towards configurations styled for unit furniture or shelf-mounting, and the case might be in the latest type of wood (oiled teak, walnut and etc.),

in metal or in a plastic material designed to look like wood.

The amplifier section has a great deal in common with the individual hi-fi transistorized amplifier. Power output can range from two times 10 W to two times 40 W, based on continuous power (e.g., r.m.s. volts across the load), and be half as much as this again when the "music power" rating is adopted. The only *real* power, of course, is that based on the square of the r.m.s. voltage across the load divided by the load resistance. One should be on the look out for so-called "peak power", which tends to give the impression that the amplifier has double the power that it really has. Based on continuous sine wave signal, the voltage across the load passes through peak twice per cycle, and to get this double-power impression the peak value of the load voltage is used in the power equation. This is $\sqrt{2}$ of the r.m.s. value, and as $\sqrt{2}$ squared is 2, the power as computed by the peak voltage is exactly doubled!

Music power is a different thing again and usually takes into account the two stereo channels running simultaneously on transient-type of "music" signal, as distinct from the continuous nature of a sine wave signal, and as this signal neatly bypasses the effect of power supply regulation and the thermal efficiency of the heat sinks of the output transistors, the power rating that it produces is above that based on r.m.s. load voltage. For example, a two times 10 W amplifier of continuous power might well have a music power rating of 30 W overall (the music power sum of the two channels).

Matching the "compact" amplifier and tuner-amplifier trend is the small, fully enclosed loudspeaker system, referred to as the *infinite-baffle* loudspeaker. Such loudspeakers can create some very nice sounds, though they do tend to suffer from bass roll-off (due to the limited enclosure size) and low efficiency. Nevertheless, the bass output can be emphasized by wall or corner-of-the-room positioning, due to acoustic l.f. magnification effects, the l.f. energy at the rear being communicated to the front, and the low efficiency is overcome by the substantially large power of contemporary transistorized amplifiers. Moreover, the small loudspeakers can be easily arranged in almost any room for the best stereo effect, and mounting the stereo pair on small shelves just above furniture height in adjacent corners of the room (on the short dimension side) solves many problems.

The larger the enclosure the better the efficiency as a general rule; also the better the bass response; but it would be undesirable to use loudspeakers of very low efficiency with a stereo amplifier having less than, say, 8 W r.m.s. power per channel. The majority of tuner-amplifiers, however, yield at least 10 W per channel (r.m.s. power).

Amplifier inputs are available for magnetic cartridge (hovering around 5 mV sensitivity—e.g., for full output power) equalized to RIAA, for an "auxiliary" source (unequalized) at about 150 mV and for a tape recorder at about 200 mV for high-level replay. Some more advanced designs have a tape head input which is also equalized to suit the tape velocity at about 4 mV, while the majority provide an output for feeding the amplifier programme signal into a tape recorder for making a tape recording.

One interesting feature about tuner-amplifiers, bearing in mind the multiplicity of circuits involved, is the relatively small size. The Fisher 160-T, for

example, rated at 20 W music power per channel (about 15 W continuous r.m.s. power) measures only $15\frac{1}{4} \times 11\frac{1}{4} \times 3\frac{1}{4}$ in. The Lux HQ555 with a 50 W rating is $17\frac{3}{4} \times 12\frac{1}{2} \times 6$ in. The best models have very low distortion figures, well within the DIN 45-500 specification, often well below 0·5 per cent. The quality potential is thus very high indeed.

An impression of overall size can be gleaned from the Heathkit (Daystrom Ltd.) Model AR17 tuner-amplifier shown in Fig. 11.1. Another example, this

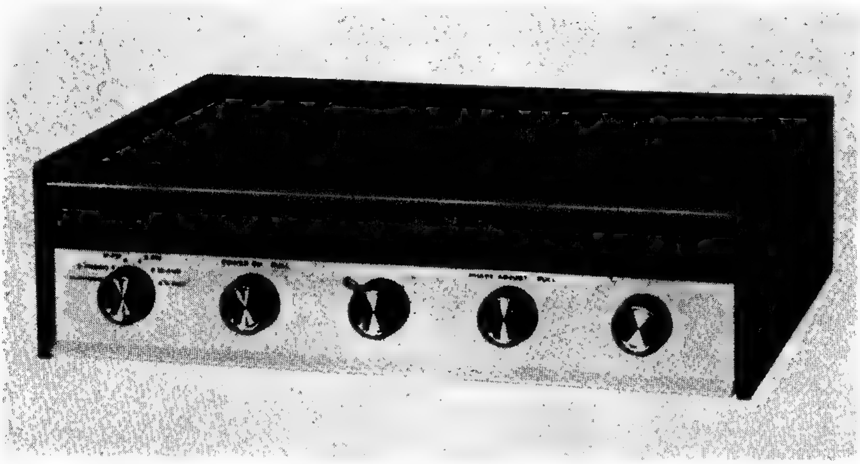


FIG. 11.1. *Tuner-amplifier, Model AR17 by Heathkit (Daystrom Ltd).*

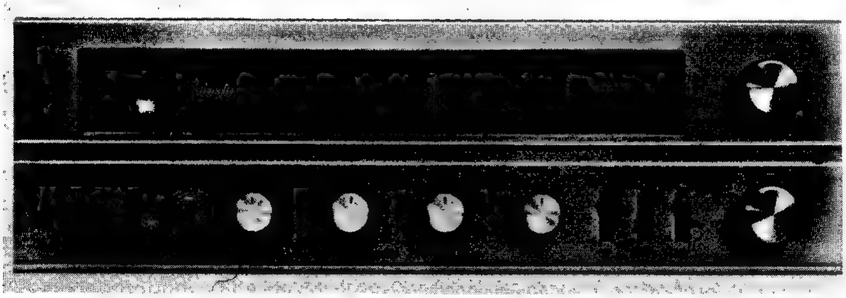


FIG. 11.2. *Tuner-amplifier by Sansui. This is an example of the very few valve models of this kind of equipment now available.*

time with valves—which must be stressed here is very unusual nowadays—is the Sansui Model 250 shown in Fig. 11.2. The vast majority of tuner-amplifiers, of course, use semiconductors, transistors of all kinds, germanium and/or silicon diodes, ceramic or crystal resonators for high selectivity, and integrated circuits. Most models are composed of printed circuit boards, and the very neat internal layout of a very up-to-date tuner-amplifier is shown in Fig. 11.3. Directly below the mains transformer on the right can be seen the i.c. i.f. strip and below this the f.m. front-end. To the left of the transformer is the stereo decoder, the remaining sections being concerned with the audio signals.

It is, of course, impossible in this book to delve into all the various aspects of hi-fi amplifiers, and readers interested in this subject might well find my companion volume, *The Hi-Fi and Tape Recorder Handbook*, of interest.

Tuners, too, are becoming highly sophisticated, and quality enthusiasts already with a good stereo amplifier will be more inclined to invest in a tuner as

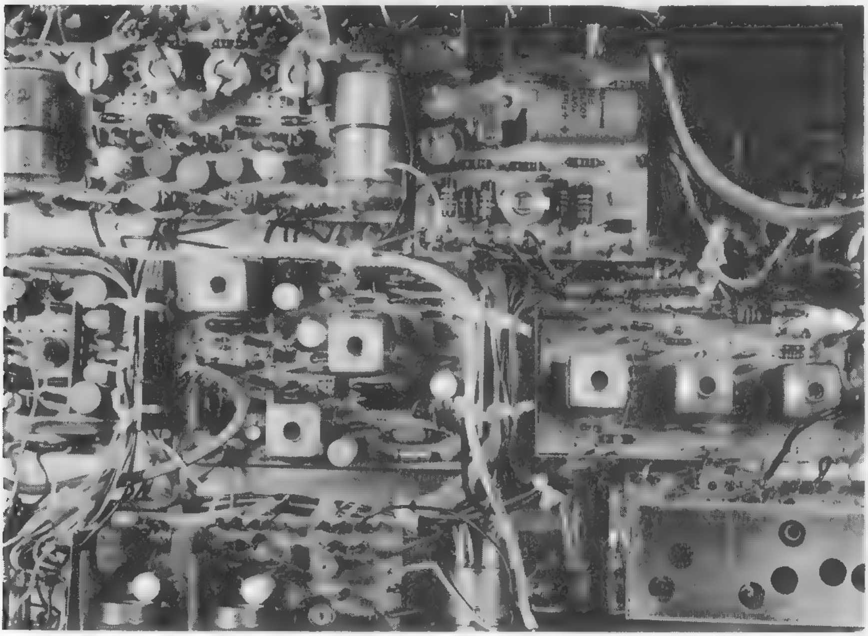


FIG. 11.3. *View inside a solid state tuner-amplifier.*

a partner. Again, the tuner may be stereo-ready or without a decoder but with facilities for easy installation of one when the time comes. Decoders are often made in the form of edge-connecting "modules" which can be plugged into a matching socket already wired in the main body of the tuner. The British Armstrong tuner adopts this technique. Past chapters have dealt with all the various aspects of design and servicing so there is no point in repeating the information here.

It is noteworthy, however, that tuners are being styled to fit in with the general idea of unit furniture and shelf-mounting hi-fi systems. Typical in this respect is the Goodmans f.m./a.m. stereo tuner, called the "Stereomax". This is shown in Fig. 11.4 partnered with the Goodmans "Maxamp" and a mini speaker from the same firm. The "Stereomax", too, has a plug-in decoder module.

One of the first British firms to go into the i.c.-designed tuner was Truvox with the FM200IC, followed by Leak (Stereofetic tuner). Some idea of the space-saving of the i.c. design is revealed in Fig. 11.5 where the three separate subchassis sections shown outside the tuner are now replaced by the single section shown fitted inside the tuner. This includes a built-in stereo decoder. The i.c.s are adjacent to the front-end section on the left, looking from the rear.

Other tuner examples are shown in Figs. 11.6 and 11.7, the very modern

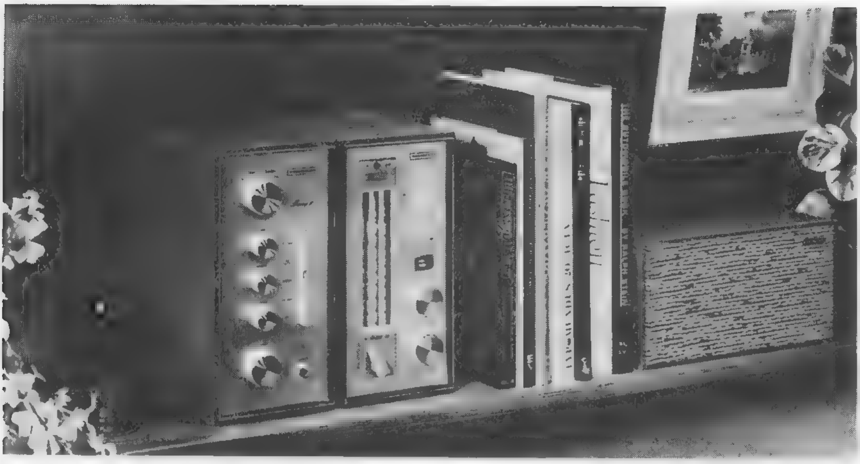


FIG. 11.4. *Shelf-mounting audio by Goodmans, showing the "Maxamp", using silicon transistors, the "Stereomax" tuner and a matching "mini" speaker.*

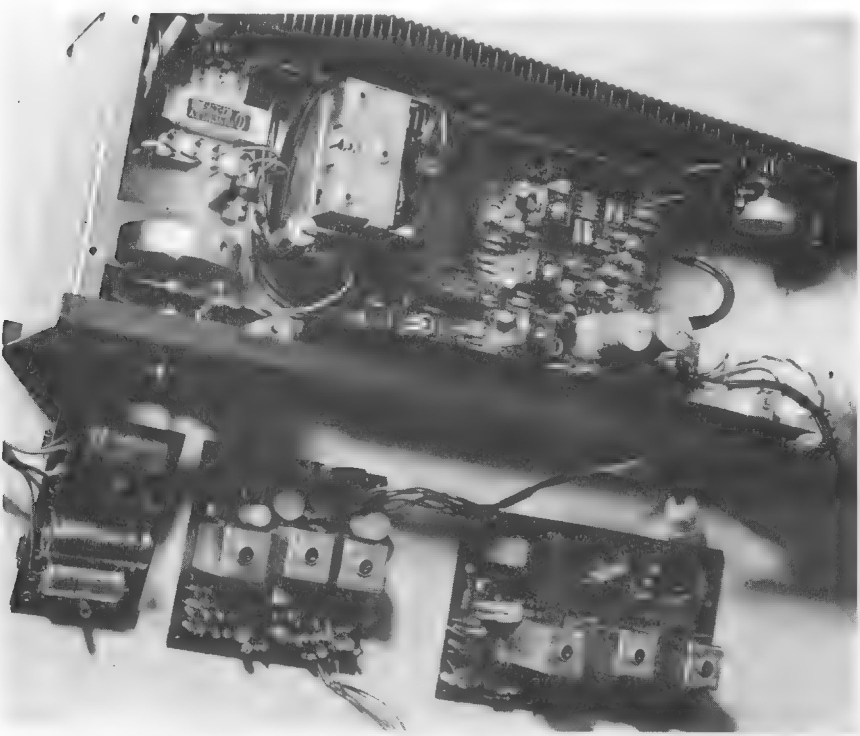


FIG. 11.5. *Truvox FM200IC stereo tuner. The panel shown within the tuner case contains the latest i.c. system of the 200-series. This replaces the three sections, shown outside, of the earlier 100-series.*

Beomaster 5000 by Bang and Olufsen and the RT40 by Grundig. The first is for f.m. only while the second has l.w., m.w. and s.w. bands as well as f.m.

TUNER SPECIFICATIONS

Limited Output

A great deal about the performance of an f.m. tuner can be gleaned from a set of curves such as shown in Fig. 11.8. The curve labelled "output" shows the performance of the limiter in filtering the a.f. output; it also shows at what aerial signal full limiting occurs, in this case about 1·5 μV, which is very good. Maximum output is given relative to 0 dB, which can be at any a.f. voltage as set by the a.f. level control or preset. The curve shows that the a.f. output is about 15 dB down when the input is 0·3 μV. The s/n curves reveal the signal/noise ratio on both mono and stereo. It will be seen that the s/n ratio on both mono and stereo becomes maximum (approaching —60 dB) with an aerial input of about 40 μV. The significant difference between the mono and stereo s/n ratios is indicated at the threshold stereo switching level, for the curve applies to a tuner in which the decoder is automatically switched on by the pilot tone (see Chapter 7). The difference is about 10 dB, which is less than the 20 dB expounded by some authorities. However, the stereo s/n ratio is closely geared to the design of the decoder and the s/n ratio pertaining to the regenerated subcarrier. It will be seen, too, that stereo operation is indicated by the pilot lamp or "beacon", as it is sometimes called, glowing.

The output curve can be obtained by feeding modulated v.h.f. signal into the aerial socket and measuring the a.f. over a range of inputs and then plotting the output against the input signal. Modulation is typically 30 per cent, though this can be at any other value provided the depth adopted is indicated, especially when absolute a.f. signal voltage is recorded as distinct from mere changes in signal level in decibels.

Signal/noise Ratio

The s/n curves are a little more difficult to obtain. The s/n ratio is fundamentally the ratio of the output power due to the signal to that due to random noise measured across the a.f. output socket. In actual fact, the ratio of signal-plus-noise *to* noise does not differ very much from the real s/n ratio when this is not less than about 20 dB, so it is often possible to calculate the s/n ratio from the expression (s + n)/n ratio. Filters, though are often incorporated to delete hum and signal components above 15 kHz.

Basic method of measurement involves feeding in a modulated v.h.f. signal (typically 30 per cent), establishing an a.f. datum and then switching the modulation off without changing the v.h.f. signal amplitude. The sensitivity of the read-out device, usually a millivoltmeter, is then substantially increased until the original datum is again reached by the meter, the number of decibels of gain so switched into the meter or read-out circuit (or, conversely, the number of decibels of attenuation switched *out*) represents the s/n ratio, expressed in decibels. This process is repeated over a wide range of v.h.f. inputs (or, at least, until the curve flattens out as at about 50 μV in Fig. 11.8) and the point-to-point s/n ratio plotted against input signal.

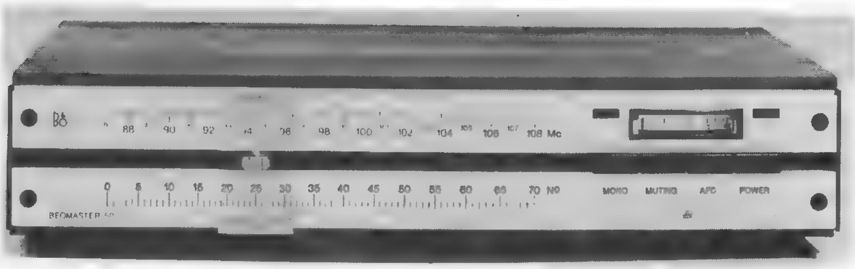


FIG. 11.6. *High quality f.m. tuner by Bang and Olufsen, the Beomaster 5000. There is a matching amplifier for this.*

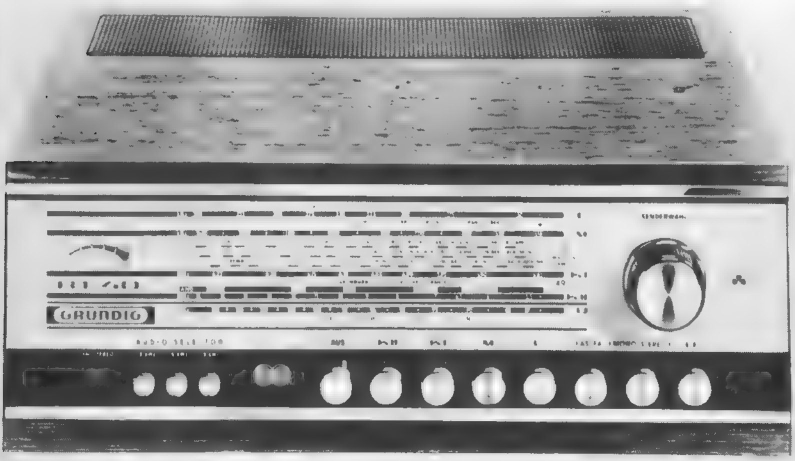


FIG. 11.7. *Model RT40 tuner by Grundig. This covers the a.m. bands as well as f.m., mono and stereo.*

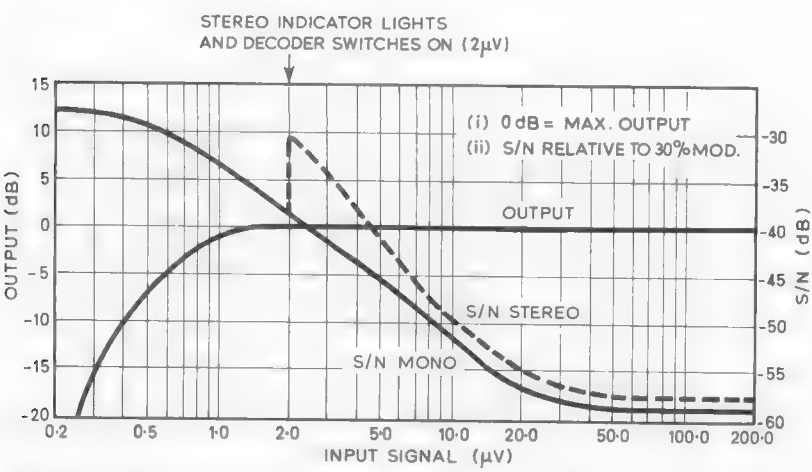


FIG. 11.8. *Curves showing the performance of an f.m. tuner. These are fully explained in the text.*

To secure the same curve for stereo the v.h.f. input needs to be a multiplex signal (from a multiplex v.h.f. generator or BBC test signal) and the s/n ratio can be plotted for each channel if necessary, though both channels are usually within a dB of each other. Another way involves taking down the multiplex v.h.f. input until the decoder *just* switches off, measuring the mono s/n ratio at that input and then adjusting the threshold preset until the decoder comes in and again measuring the s/n ratio. The latter, of course, will always be worse than the former owing to the effect of the pilot tone and the decoding process. This scheme yields comparative s/n ratios (mono/stereo) at the decoder threshold point, which is often an important consideration in weak signal areas.

Sensitivity

The specification *usable sensitivity* would, for quality tuners, refer to the v.h.f. input amplitude which provides approaching full limited output at a s/n ratio not less than —40 dB. In Fig. 11.8 this would be about 2·5 μV, though the limiting would have been completed when the input reached about 1·5 μV, as we have already seen.

The usable sensitivity on stereo is always less than on mono; in Fig. 11.8 —40 dB s/n ratio is given with a stereo input of about 4·5 μV, as against 2·5 μV for mono.

Some very high quality tuners have mono s/n ratios up to about —65 dB, with stereo almost the same but possibly a little less good, depending on the design of the decoder.

Capture Ratio

This refers to how far below the amplitude of a wanted f.m. signal an unwanted signal in the same channel must be before it is not subjectively "disturbing". The ratio is given in decibels, and for top-flight tuners the ratio might be as small as 1·5 dB. This means that provided the wanted f.m. signal is about 20 per cent above the level of the unwanted signal (in terms of signal *voltage* at the aerial socket) the unwanted signal will not disturb the reception unduly. The "disturbance value" appears to differ between authorities, but it is typically—30 dB. For hi-fi reception, though, the ratio would have to be at least 46 dB, or better. The ratio improves, of course, as the wanted signal reduces in strength.

We saw in an earlier chapter that the *capture effect* relates to the ability of an f.m. tuner to suppress an unwanted transmission even when it is on the same frequency as a wanted transmission, irrespective of the type of modulation carried by the offender. The effect also comes into play when the unwanted transmission is removed from the frequency of the wanted transmission yet still remaining in the receiver's pass-band. How well the capture effect works depends on the excellence or otherwise of the design of the i.f. channel and the f.m. detector or discriminator, and on the alignment of these circuits.

When an in-channel or adjacent channel unwanted transmission is stereo-encoded, as also the wanted transmission, interference can sometimes occur due to the subchannel modulation components and give rise to "birdies" or similar noises as explained in Chapter 7. The alignment and design of the f.m. detector or discriminator have a bearing on this, as too does the degree of

filtering between the f.m. detector/discriminator and the decoder input.

Selectivity

The selectivity of a tuner represents its ability to separate a wanted signal from an unwanted one on a near frequency. Measurements are usually made with two signals fed into the aerial socket by way of a matching "star" network to represent the wanted and unwanted transmissions. Initially the wanted signal is modulated to 30 per cent to establish a datum but then the wanted signal is unmodulated during the presence of the unwanted one which should be modulated to 30 per cent. The ratio of the output power caused by the unwanted signal to the output power initially established by the wanted signal (when it was modulated), expressed in decibels, represents the selectivity ratio. It is usual to take readings over a range of inputs and signal spacings to establish curves with the level of the wanted signal as the parameter.

Adjacent-channel Response Ratio

This is taken in a similar way to selectivity, but the wanted and unwanted signals are spaced by 200 kHz (e.g., standard f.m. channel). Curves can be constructed over a range of input levels. Both selectivity and adjacent-channel response ratios may be taken relative to a "disturbance value" which, as for the capture ratio, might be —30 dB. Thus the dB difference between the two signals is arranged to relate to —30 dB on the test rig, the difference in decibels to this "datum" then representing the required ratio. Very good tuners can have a ratio as high as —80 dB.

Frequency Response

This is the a.f. output against a decibel scale over a range of modulation frequencies applied to the v.h.f. carrier fed into the aerial circuit. The v.h.f. signal level should take the set into full limiting and maximum s/n ratio. A good hi-fi tuner should be within $\pm$ 3 dB from 20 Hz to 15 kHz, but account has to be taken of the inbuilt de-emphasis if the f.m. signal fails to carry pre-emphasis.

The frequency response should be taken on both stereo channels when a multiplex generator is available.

Rejection Ratios

The output at a.f. due to spurious acceptance channels in the tuner relative to the a.f. output at a datum level due to a signal in the normal channel is expressed in decibels. Common spurious channels and their dB ratios are image 80 dB, i.f. 90 dB, spurious generally 90 dB and a.m. supression 50 dB. The last one is closely controlled by the design and alignment of the f.m. detector/discriminator and on the amplitude limiter.

Muting Level

This refers to the v.h.f. signal input level that is necessary to lift off the inter-station muting. It is often adjustable by a threshold control (preset) and if usually set to about 5 μV on good tuners.

Distortion

The total harmonic distortion (t.h.d.) due to modulation all the way from the

aerial to the a.f. outputs should not be more than about 0·5 per cent for hi-fi applications. It is usually higher on stereo due to the action of the decoder and the distortion on the multiplex signal itself; but a good tuner will have about 0·5 per cent stereo and 0·3 per cent mono.

Crosstalk

This should not be worse than 20 dB at 1 kHz (it is usually much better than this on good hi-fi tuners, approaching 40 dB), though it may fall a little at spectrum extremes. Loss of stereo will result from poor crosstalk performance between about 500 Hz and 5 kHz. Details of measurement using the BBC signal are given in Chapter 7.

Decoder Switching

Tuners incorporating auto mono/stereo switching (also described in Chapter 7) usually have a threshold control and a common setting is for the switching to occur at a v.h.f. (multiplex) input of 5 μV.

SCA Suppression

Tuners with SCA (see Chapter 7) filters should reject the SCA frequency to the extent of about 50 dB.

Subchannel Suppression

Subchannel components at 19 kHz and 38 kHz from stereo tuners are often filtered out by rejectors in addition to the normal roll-off effect of the de-emphasis. Such rejection is essential for recording stereo radio when the tape recorder bias oscillator signal beats against them (see Chapter 7). Good tuners are already equipped with such filters giving a suppression ratio of about 50 dB. The filtering may in part incorporate the de-emphasis.

Quieting

This term relates to the s/n ratio already dealt with, but it may be applied differently in that decibels of "quieting" are quoted to indicate noise reduction between the noise at the a.f. output minus a v.h.f. signal at the input and that when a v.h.f. signal is applied. Some makers give a μV aerial input required for 30 dB quieting. The s/n ratio appraisal is desirable, however.

Quietening

This term, not uncommonly confused with quieting (even by some makers!), is another way of expressing the interstation muting.

Din 45-500

This is a set of standards composed by the German Deutscher Industrie Normenausschuss to establish *basic minimum* requirements for hi-fi equipment. Although it falls below the standards considered desirable for hi-fi by some authorities it is, nevertheless, a commendable piece of work, since it does at least give some basis from which to judge equipment. It is used extensively by audio equipment makers in Europe and is referred to in their specifications and service manuals; but, as in Great Pritain, quite a lot of equipment imported from overseas possesses specifications far in excess of the basic DIN minimum.

Nearest British documents to the DIN are (i) that evolved in 1960 by the Audio Group of the British Radio Equipment Manufacturers' Association (BREMA) and finally published as British Standard 3860:1965 and (ii) British Standard 4054: 1966. Both of these however, are mainly concerned with measuring and expressing the performance of (i) audio amplifiers and (ii) receivers for a.m. and f.m. Some of the above specifications are based on BS 4054:1966.

Extracts translated from the DIN section dealing with the minimum requirements for f.m. receivers, including tuners, are given below.

Range of Reproduction

To cover from at least 40 Hz to 12·5 kHz with permissible deviations relative to 1 kHz of ± 3 dB from 40 to 50 Hz, ± 1·5 dB from 50 Hz to 6·3 kHz and ± 3 dB from 6·3 to 12·5 kHz.

Differences between Stereo Channels

Maximum deviation between the channels is 3 dB within the range 250 Hz to 6·3 kHz.

Distortion Factor

Non-linear harmonic distortion to be equal to or better than 2 per cent at 1 kHz with a deviation of 40 kHz. Stereo channels to be both within this specification.

Stereo Separation

To be equal to or better than 26 dB from 250 Hz to 6·3 kHz and 15 dB from 6·3 to 12·5 kHz.

Signal/noise Ratio (unweighted)

To be equal to or better than 46 dB, mono or stereo, from 40 Hz to 15 kHz.

Pilot Tone Signal/noise Ratio (unweighted)

To be equal to or better than 20 dB at 19 kHz and 30 dB at 38 kHz.

Characteristics to be Stated in Specifications

(i) Recommended aerial impedance.
(ii) Aerial signal voltage at the recommended impedance.
(iii) Lower impedance and permitted loading at the l.f. output.
(iv) Output a.f. voltage at a modulation depth (deviation) of 40 kHz.

Recommendations for Connectors

DIN "standard" socket (see Fig. 11.9) to be employed, with contact No. 3 for the left (A) channel and contact No. 5 for the right (B) channel.

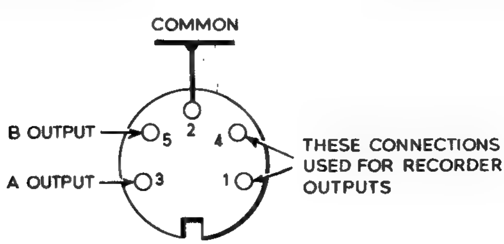


FIG. 11.9. *The DIN-standard socket. Connectors 3 and 5 are used for tuner A and B outputs.*

CRYSTAL GAZING

What of the future? Well, there is little doubt that the high quality tuner-amplifier is going to become even more popular, and it will probably represent the radio receiving equipment in many homes, taking over from the old-style radio and radiogram. Popularity is likely to be shared with the so-called "music compacts". This type of domestic equipment is basically a plinth-mounted record player of size suitable for shelf or unit furniture mounting containing a good quality stereo amplifier. Input facilities are available for radio and tape recorder signals, though some models incorporate a radio tuner giving mono and stereo f.m. and not uncommonly a.m. reception, too. The loudspeakers are separate, essential for the best stereo effect and reproduction, but all that is involved in getting this sort of equipment operational is plugging into the mains supply and connecting the loudspeakers, which are invariably infinite-baffle type.

The thermionic valve is rapidly becoming a device of the past in almost all aspects of electronics, domestic and commercial, and although the transistor is in full flight at this time, integrated circuits are already penetrating the price barrier and becoming reasonable propositions in high quality audio equipment as we have seen in some of the earlier chapters. Indeed, right now it is possible to produce a complete stereo tuner-amplifier from monolithic i.c.s, for i.c.s are available for i.f. amplification and f.m. detection, for the multiplex decoder, the audio preamplifiers and the main amplifiers giving up to 5 W audio per channel. Up to 15 W per channel can be obtained from a hybrid thick-film main amplifier system!

Another tuner and tuner-amplifier feature of the future is the ceramic or crystal resonator in the f.m. i.f. channel. Such a device significantly aids selectivity without impairing the required bandwidth and hence makes it possible to delete all traces of adjacent-channel interference, especially desirable on stereo transmissions. Already, imported and British equipment is beginning to adopt the resonator in conjunction with ordinary tuned i.f. filters.

The decoder, incidentally, is tailored from a single dual-in-line flatpack i.c. (by Motorola) which, to operate, requires three external tuned circuits and one or two resistors and capacitors. The i.c. also provides an output of 12 V, 40 mA in the presence of pilot tone for a stereo indicator lamp. Stereo separation is as high as 34 dB while all that is needed to get the A and B outputs is a multiplex input from an f.m. detector of about 200 mV r.m.s. ... So to the future.

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F.M. Radio Servicing Handbook 2nd Edition

Gordon J. King, Assoc. IERE, MIPRE, MRTS

The introduction of very-high-frequency broadcasting services has brought into use the v.h.f. radio receiver, which may largely displace the older type of receiver in many areas. These v.h.f. transmissions employ frequency modulation (f.m.) as distinct from amplitude modulation (a.m.), which has been used for earlier broadcasting services, and these receivers are designed to accept transmissions thus modulated. Every service engineer should therefore understand the nature and principles of f.m. operation, and obtain practical knowledge of the adjustment and servicing procedures involved, so that he will have no difficulty in dealing with f.m. receivers.

This handbook has been written by an experienced radio engineer with the aim of providing this theoretical and practical knowledge in a form helpful to all concerned with service work. The book is intended not only for professional service engineers, however, but also for amateur enthusiasts interested in the construction of f.m. equipment and for radio students. The style is straightforward and, as far as possible, non-mathematical.

The theory of the subject is first discussed in some detail, and the particular advantages of f.m. in alleviating interference and improving audio quality are considered. Circuit arrangements are then discussed stage by stage, while chapters are devoted to combined a.m./f.m. receivers and tuners, to f.m. radio stereophony, and to aerial problems. The alignment and servicing of f.m. equipment is described; the final chapter deals with the specifications used for f.m. high-fidelity systems for domestic use.

F.M. Radio Servicing Handbook is an important part of the Gordon J. King series of servicing guides. These guides form an authoritative and up-to-date reference library, invaluable to the busy service engineer and the keen radio experimenter.

A list of other books by Gordon J. King faces the title page.

The Author

Gordon J. King is the holder of several patents and created the low-noise transistor v.h.f. and u.h.f. 'Telebooster' and the powered set-top aerial. He was one of the initiators of high quality sound reproduction before the term Hi-Fi was in vogue and this is still one of his chief interests. He is a well known reviewer of hi-fi equipment and the author of at least fifteen books, from radio to colour television and from audio to hi-fi and tape recorders, covering the whole range of his interests. A Member of the Incorporated Practitioners in Radio and Electronics and of the Royal Television Society, he is also an Associate of the Institution of Electronic and Radio Engineers. He is a graduate member of the Institution of Technical Authors and Illustrators and holds a diploma in television engineering.

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F.M. Radio Servicing Handbook

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